WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

FEBRUARY 1950

VOL. XXVII TWO SHILLINGS AND SIXPENCE No. 317

CARPENTER
POLARIZED RELAYS

TMC

Chief Characteristics*

High operational speeds • Freedom from contact rebound and positional error • Good contact pressure • Accuracy of signal repetition • Sensitivity of the order of 0.12 Mw. D.C. • Ease of contact adjustment • Contact gap a function of input power.

Due to the unique combination of characteristics* inherent in the design of Carpenter Polarized Relays, they are providing practical solutions to many problems in scientific research and electrical engineering.

Their high standard of performance makes them suitable for use where signal impulses of varying time duration must be repeated with the utmost accuracy, *for instance in* telegraphy, measurement, protection and tele-control schemes.

The relay operates successfully direct from valves, the barrier layer type of photo cell, and also from thermo-couples. In the amplification of very weak D.C signals, the relay's almost perfect contact performance makes it suitable for the inversion of the D.C. into A.C. to permit electronic amplification.

Three Main Types are available.

TYPE 3

has maximum speed : maximum contact pressure and sensitivity : minimum transit time.
Dimensions **: 120mm x 61mm x 33mm Weight 625 gm.**

TYPE 4

has longer contact travel than the Type 3, with medium speed and sensitivity. Also available in double-pole, double throw version with self-synchronising contacts.
Dimensions **: 91mm x 56mm x 25mm. Weight 370 gm.**

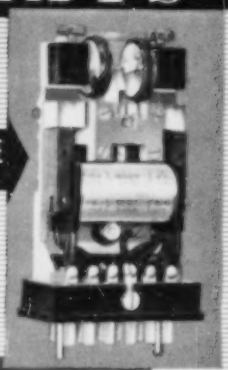
TYPE 5

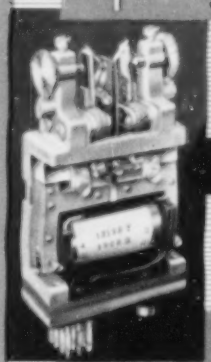
miniature relay with outstanding performance for size. Rugged design of exceptional thermal stability.
Dimensions **: 56mm x 37mm x 19mm. Weight 140 gm.**

Note : All dimensions quoted exclude connection terminals.

Complete details and performance data will be willingly supplied on request to the Manufacturers.

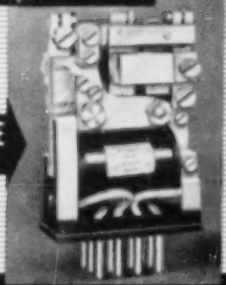
TYPE 3





TYPE 4

TYPE 5



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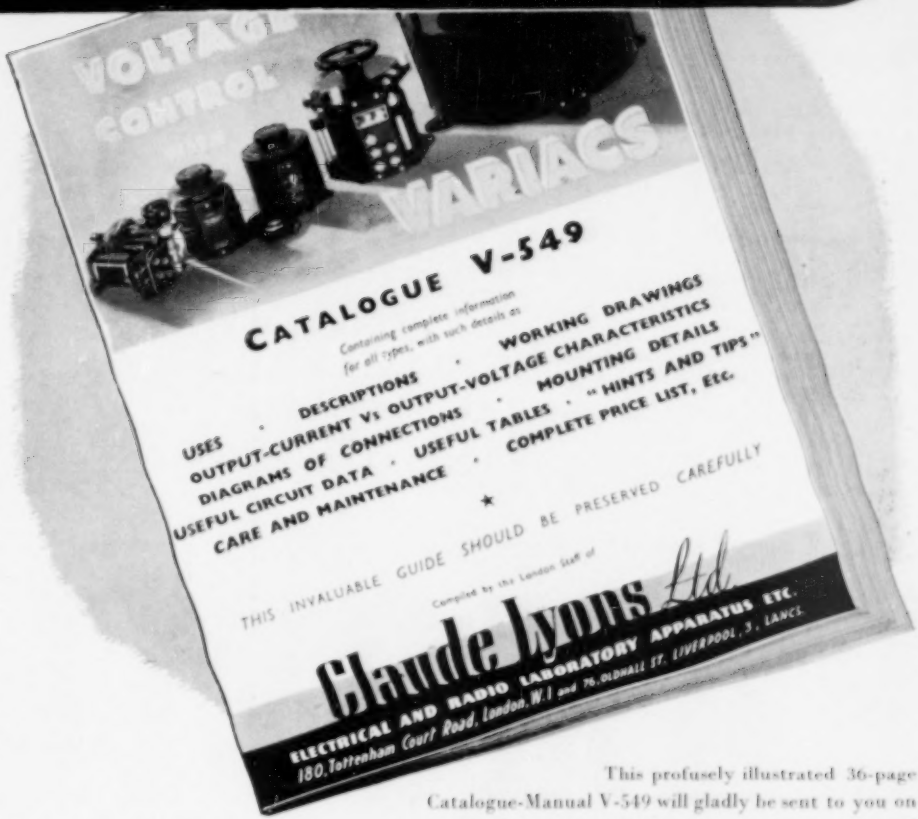
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VARIACS
CATALOGUE V-549
Containing complete information for all types, with such details as:
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OUTPUT-CURRENT Vs OUTPUT-VOLTAGE CHARACTERISTICS
DIAGRAMS OF CONNECTIONS · MOUNTING DETAILS
USEFUL CIRCUIT DATA · USEFUL TABLES · "HINTS AND TIPS"
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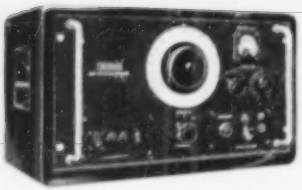

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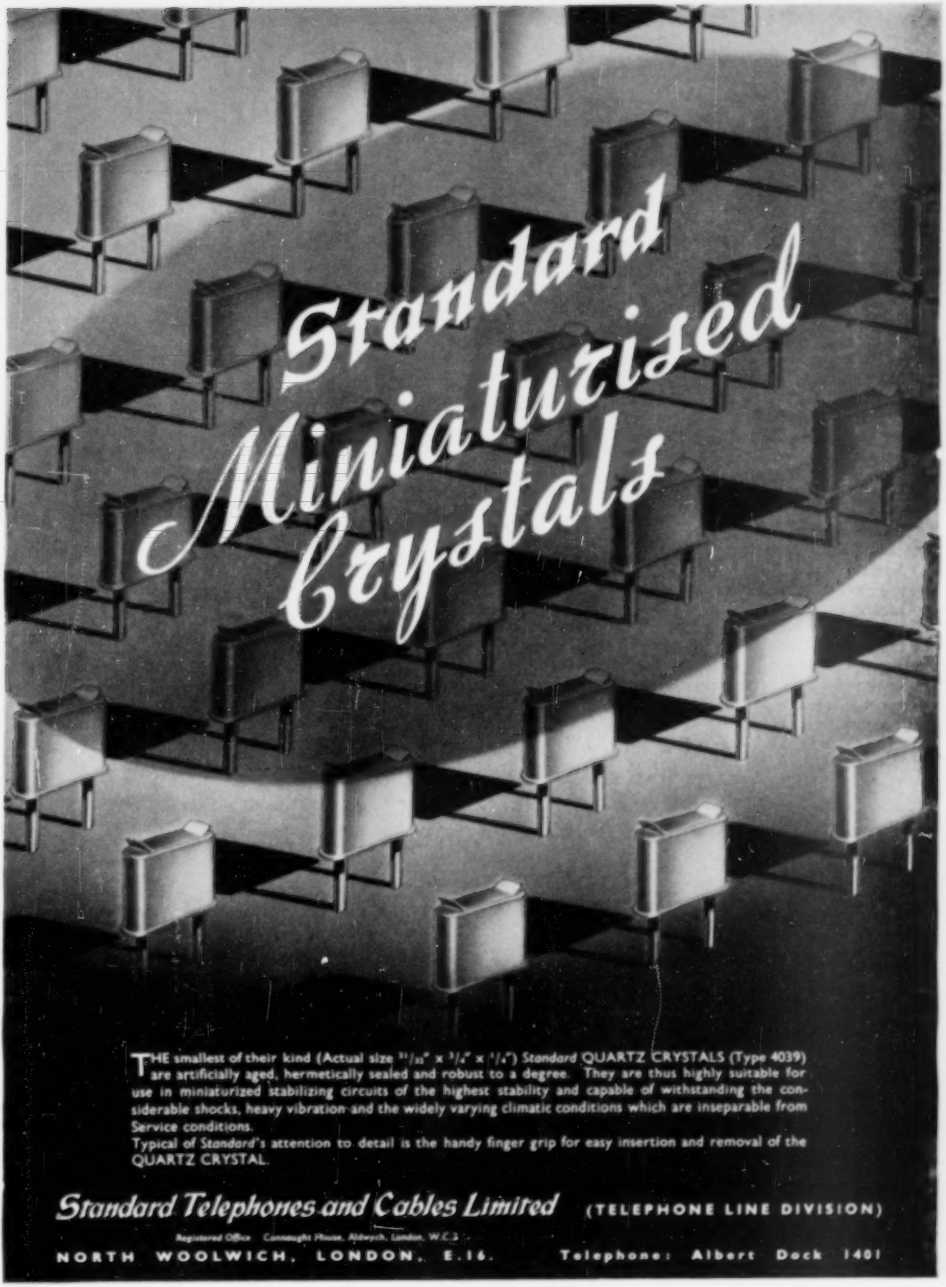

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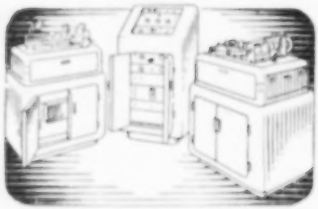

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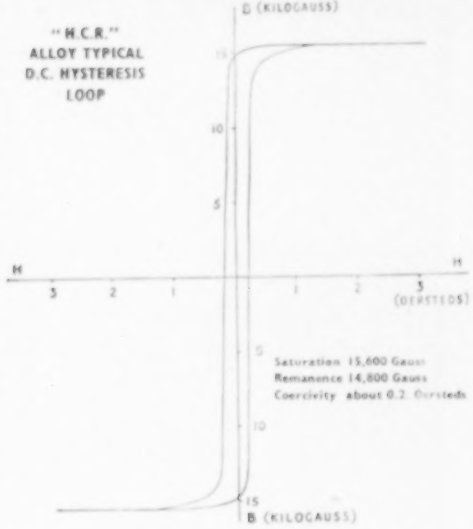

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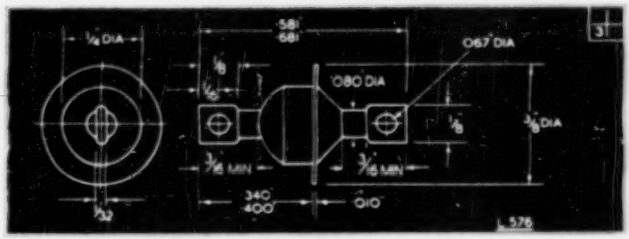

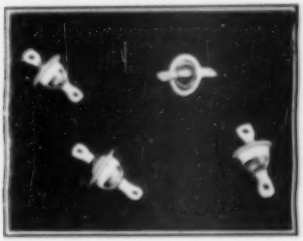

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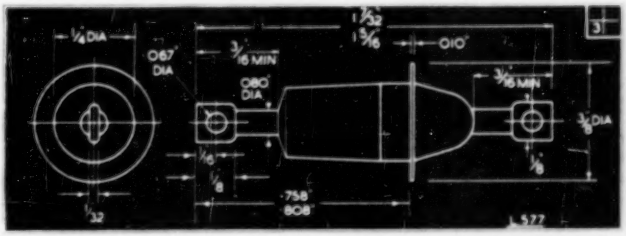
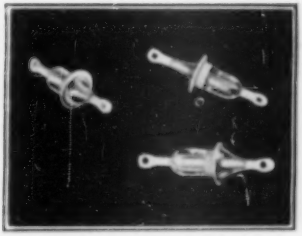
U.K. Patent applied for.

L.576
750 v. DC. Working at 40,000 feet.
1,500 v. DC. Working at sea level.

These hermetically sealed terminals employing glass to metal seals, are made for the sealing of transformers or other components. They are of course, also useful as insulating pillars where high resistances are required.





L.577
1,500 v. DC. Working at 40,000 feet.
3,000 v. DC. Working at sea level.

They withstand instantaneous and repeated thermal shocks of at least 250° C., and will support at least 40 lbs. per sq. in. air pressure without leakage.
These terminals are supplied tin-plated to permit soldering with modern resin cored solders, or solder pastes.
Further details and prices on application.

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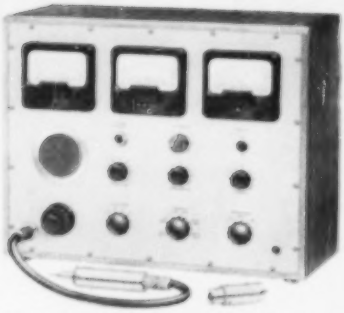
Type 2002

is a selective tube voltmeter for the measurement of a.c. voltages in the high frequency range. The instrument is particularly designed for use in radio laboratories for measurements on radio receivers, Radar IF-circuits, for the control of signal generators, co-axial carrier frequency systems, etc. The high sensitivity makes the instrument very suitable for measuring small a.c. voltages, for example from aerials. The instrument has a meter for modulation measurements.



Frequency Range : 20kc/s—27 Mc/s in 4 ranges.
Voltage Range : Full scale deflection for 10—100—1,000 microvolts and for 10 and 100 mV. By using the Attenuator 2002a 1 and 10 volts.
Input Impedance : 5 megohms in parallel with 12 pF.
Intermediate Frequency : 1,650 kc/s.
Bandwidth : —3 db at ± 3.5 kc/s, — 10 db at ± 6kc/s,—40 db at ± 13 kc/s,—80 db at ± 25 kc/s.
Frequency Accuracy : About 2% ± 2 kc/s.
Voltage Accuracy : ± 0.5 db in the broadcast frequency range and ± 1 db in the short-wave range.
Reference Voltage : An oscillator delivering exactly 1 mV, 100 kc/s is built-in.
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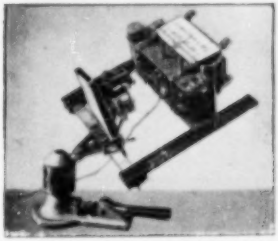

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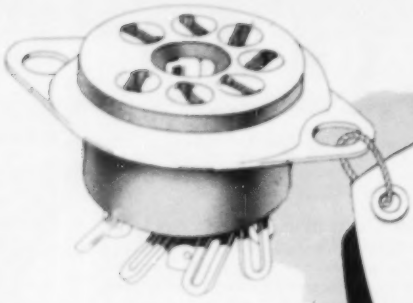

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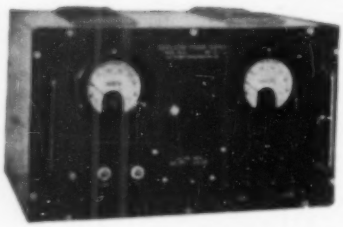

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Stability : Better than 0.1%.
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Output Ripple : Less than 2mV. R.M.S.
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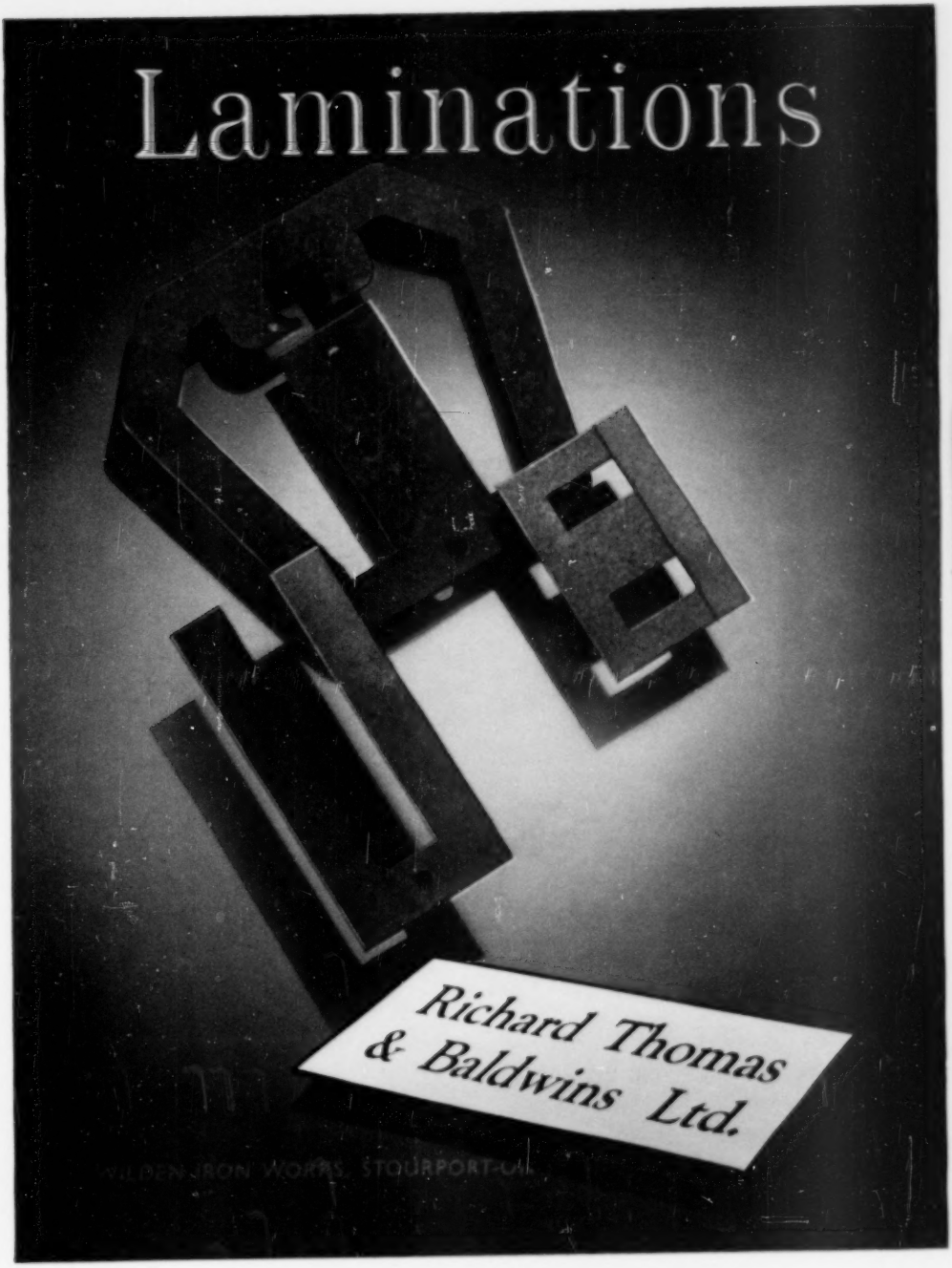
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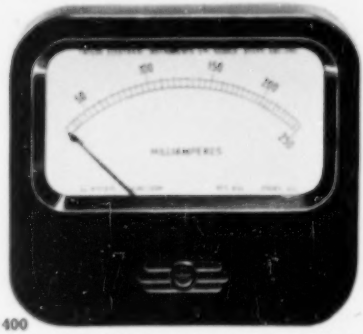

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Axial	2.4" (6 cm.)
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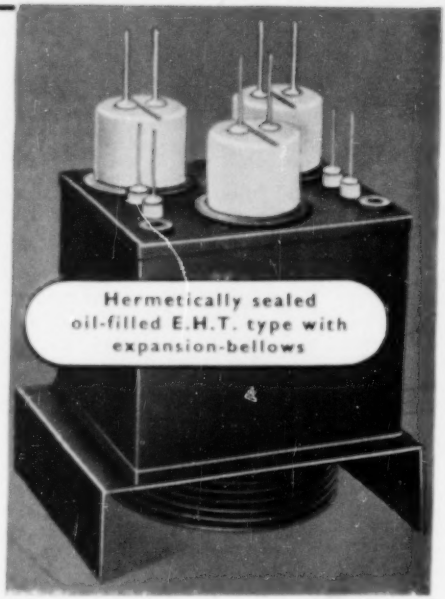

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Approximate Arc Voltage Drop (volts)	9.0
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BASE:— IO8

LIST PRICE:— £2.12.6 (Provisional)

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Maximum Overall Length (mm)	110
Maximum Diameter (mm)	40
Maximum Seated Height (mm)	97

All maximum ratings are absolute values, not design centres.

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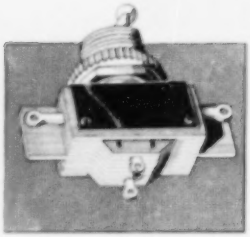

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FEBRUARY 1950

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DF72	Sharp cut-off R.F. Pentode	41.2	1.25	25	67.5	67.5	0	1.7	1.0	650	—
DF73	Variable-mu R.F. Pentode	41.2	1.25	25	67.5	67.5	0	1.7	0.8	450	—
DAF70	Single Diode A.F. Pentode	41.2	1.25	25	67.5	67.5	0	0.9	0.45	300	—
DL72	A.F. Output Pentode	41.2	1.25	25	90	90	−3	1.3	0.67	500	0.047
EF70	High Slope R.F. Pentode with short g3 base	38	6.3	200	100	100	−2	3.0	2.3	100	—
EF72	High Slope R.F. Pentode	38	6.3	150	100	100	−1.4	7.0	5.0	200	—
EF73	High Slope non-R.F. Pentode	38	6.3	200	100	100	−2.0	7.5	5.0	250	—
EC70	R.F. Triode for use as Oscillator up to 500 Mc/s	38	6.3	150	100	—	−2.0	13	5.5	3.6	0.75 500 Mc/s
*EA76	Single Diode	25.4	6.3	150	150 (r.m.s.)	—	—	9 (max)	—	—	—
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WIRELESS ENGINEER

Vol. XXVII FEBRUARY 1950 No. 317

Stresses in Magnetic Fields

THIS is a subject that we considered in the Editorial of December 1946, in which we discussed the lack of agreement on the magnitude of the various stresses. We propose now to draw attention to several points which put the matter beyond any doubt, and entirely remove the vagueness that seemed to envelop the subject.

If a uniformly-wound toroid of wood of cross-section A is split across a diameter, and the two halves pulled apart a short distance δs, the current being adjusted to keep the flux constant, there is no change in H or B in the wood, but there are now the air-gaps of total volume $2A.\delta s$ with their magnetic energy of $H^2/(8\pi)$ ergs per cubic cm. As this energy must be equal to the work done in pulling the two halves apart, the force exerted must be $2AH^2/(8\pi)$, that is $H^2/(8\pi)$ dynes per cm^2; we use c.g.s.e.m. units in which B and H in air are numerically equal. The same would be true without the wooden ring, which is merely a support for the winding.

How and where is this pull exerted by the field? On the turns of wire which carry current in the magnetic field due to all the other turns. Every part of a turn experiences a force tending to increase the size of the turn, but as the magnetic field is stronger on the inside of the ring because of the shorter path, there is a resultant force tending to move the turn towards the centre of the ring.

If we now introduce an iron ring instead of the wooden one, with the same initial current as before, there will be no change in H in the material of the ring, and no change in the magnetic field in which the winding is situated, and consequently no change in the forces on the winding, or the resultant catapult action of the winding in opposing the applied pull. This must still be equal to $H^2/(8\pi)$ dynes per cm^2 of cross-section. But from the energy now stored in the air-gap we find that the total pull must equal $B^2/(8\pi)$ dynes per cm^2; hence the forces on the iron must account for the difference $B^2(1-1/\mu^2)/(8\pi)$ or $BH(\mu-1/\mu)/(8\pi)$. As before, we assume that, as the small air-gap is made, the current is increased so as to keep the flux, and therefore B and H in the iron constant. The field in which the winding is situated will also be unchanged. In the gap $H_g = B$.

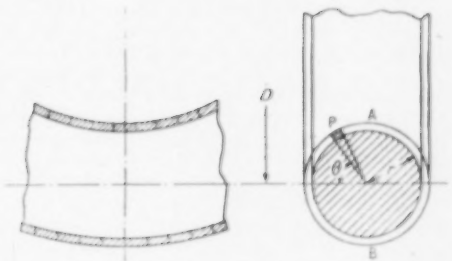


Fig. 1

Returning to the wooden ring, we will now calculate the forces on the winding in the ordinary way, in order to confirm that it gives the same result as that obtained from the energy in the air-gap.

Fig. 1 shows a section of a part of the ring uniformly wound with rectangular strip, which for simplicity of calculation is assumed to be in close contact. The value of H falls to zero over the thickness of the strip, so that the mean

value will be half that at the inner surface. At A the length of the magnetic path is $\pi(D - 2r)$ whereas at B it is $\pi(D + 2r)$. At P the mean value of H will be

$$\frac{\text{m.m.f.}}{2\pi(D - 2r \sin\theta)}$$

and the radial force on the piece of wire $r d\theta$ at P carrying a current I will be

$$\frac{I}{10} \times \frac{\text{m.m.f.}}{2\pi(D - 2r \sin\theta)} \times r d\theta.$$

Its component parallel to AB will be this multiplied by $\sin\theta$, and on integrating this over the whole turn, we have

$$\frac{I}{10} \times \frac{\text{m.m.f.}}{2\pi D} \times r \int_0^{2\pi} \frac{\sin\theta \, d\theta}{1 - \frac{2r}{D}\sin\theta}$$

$$= \frac{Ir}{20\pi D} \times \text{m.m.f.} \times \frac{2\pi r}{D} \text{ approximately,}$$

since $2r/D$ is assumed to be small.

The number of turns per cm of mean periphery is $T/(\pi D)$, and the resultant inward force per cm of periphery will be

$$\frac{IT}{\pi D} \times \frac{r}{20\pi D} \times \frac{4\pi IT}{10} \times \frac{2\pi r}{D} = \frac{4}{100} \cdot \frac{I^2T^2r^2}{D^3}.$$

The total upward force on the lower half of the ring will be D times this, that is,

$$\frac{4}{100} \cdot \frac{I^2T^2r^2}{D^2} \text{ and putting } H = \frac{4\pi}{10} \cdot \frac{IT}{\pi D}$$

this becomes

$H^2r^2/4$ dynes.

Dividing this by $2A(=2\pi r^2)$ we get $H^2/(8\pi)$ dynes for the force per cm² of cross-section.

The same result can be obtained more simply by assuming the toroid to have a square cross-section. In the integration we made an approximation, but it should be noted that the value of B in the air-gap varies from point to point, and the usual formula $B^2/(8\pi)$ is an approximation to the mean square. By assuming a uniformly-wound toroid the nature of the field is accurately prescribed and one is not troubled with magnetic leakage.

Divergent Views

In the December 1946 Editorial we pictured the lines of force crossing each square cm of the gap, splitting up into two components H and $4\pi J$, the former of which passes through the surface into the iron while the latter ends on the surface polarity. In all these formulae H is the magnetic force in the iron and not in the gap. The polarity or intensity J per unit area is equal to $(B - H)/(4\pi)$.

It is in calculating the mechanical force on J that the uncertainty arises. Is the effective field strength to be employed equal to $2\pi J$ or to $H + 2\pi J$; that is, does the H that passes through the surface into the iron play any part in exerting force on the surface polarity? If not, then the force per cm² on the polar surface is equal to

$$\frac{B - H}{4\pi} \times \frac{B - H}{2} = \frac{B^2}{8\pi} - \frac{BH}{4\pi} + \frac{H^2}{8\pi}$$

$$= \frac{B^2}{8\pi}\left(1 - \frac{2}{\mu} + \frac{1}{\mu^2}\right)$$

which was the result arrived at by Maxwell. Since the total tension per cm² is $B^2/(8\pi)$, the difference must be the tension in the field which has passed into the iron, and this must be equal to

$$\frac{B^2}{8\pi}\left(\frac{2}{\mu} - \frac{1}{\mu^2}\right)$$

We have now seen that a part of this [viz., $H^2/(8\pi)$] represents the force on the windings. This leaves

$$\frac{B^2}{8\pi}\left(\frac{2}{\mu} - \frac{2}{\mu^2}\right) = \frac{BH}{4\pi}\left(1 - \frac{1}{\mu}\right)$$

for the part of the air-gap force per cm² which acts throughout the iron with a sort of catapult action.

The total force is made up of three components, the direct pull on the pole-face, the tension of the field which is converted into lateral forces acting on the iron, and the tension of the field which is converted into lateral forces on the windings. Generally speaking, only the first is of any practical importance. Maxwell's view that the forces in the iron may be regarded as made up of a uniform pressure of $H^2/(8\pi)$ in every direction (not only transverse to the field) and a tension of $BH/(4\pi)$ in the direction of the field, gives a resultant tension along the field equal to the difference, that is,

$$\frac{B^2}{8\pi}\left(\frac{2}{\mu} - \frac{1}{\mu^2}\right)$$

which agrees with the result obtained above.

In the 1946 Editorial we referred to the fact that Helmholtz disagreed with Maxwell, and that a different result was obtained by Moullin by assuming that the effective field strength acting on the polarity of the pole-face is not $2\pi J$ but $H + 2\pi J$. This seems more probable, for it is difficult to see why the field H should not exert any force on the surface polarity. On this assumption the force on 1 cm² would be

$$J(H + 2\pi J) = \frac{B - H}{4\pi}\left(H + \frac{B - H}{2}\right)$$

$$= \frac{B^2 - H^2}{8\pi} = \frac{B^2}{8\pi}\left(1 - \frac{1}{\mu^2}\right)$$

Since the total force is $B^2/(8\pi)$ and the force

to be transmitted to the winding is $H^2/(8\pi)$, this is undoubtedly the total force acting on the iron. Hence the assumption made by Moullin gives the total lifting force exerted on the iron, as distinct from that on the winding, and it naturally vanishes when $\mu = 1$. If this assumption were correct, the total force would only be divided into two components, that on the pole-face and that transmitted through the iron to the winding. We show later that neither Maxwell nor Moullin make the correct assumption.

We shall now consider an iron ring of square cross-section uniformly wound: if the side of the square is s and the inner, outer and mean diameters are D_i, D_o, and D respectively, and H_i, H_o and H the corresponding values of the magnetic force, then assuming with Maxwell that there is a lateral pressure of $H^2/(8\pi)$, the pressure at the inner surface on an element subtending an angle $d\theta$, and therefore having an area of $(D_i/2)d\theta \times s$, will be

$$\frac{H_i^2}{8\pi} \times \frac{D_i}{2} \cdot s \cdot d\theta$$

The pressure at the outer surface will be

$$\frac{H_o^2}{8\pi} \times \frac{D_o}{2} \cdot s \cdot d\theta$$

The resultant force on the element will act towards the centre of the ring and will be equal to the difference

$$(H_i^2 D_i - H_o^2 D_o) \frac{s}{16\pi} \cdot d\theta$$

For the total radial force on the half ring we put $\int d\theta = \pi$, and for the resultant upward component we must multiply by $2/\pi$; this gives for the total upward force

$$(H_i^2 D_i - H_o^2 D_o) \frac{s}{8\pi}$$

which, since $H_i D_i = H_o D_o = HD$,

$$\frac{H_i}{H} = \frac{D+s}{D} = 1 + \frac{s}{D} \text{ and}$$

$$\frac{H_o}{H} = \frac{D-s}{D} = 1 - \frac{s}{D}$$

is equal to $H^2 D\left[\left(1+\frac{s}{D}\right) - \left(1-\frac{s}{D}\right)\right]\frac{s}{8\pi}$

that is $\frac{H^2 \cdot 2s^2}{8\pi}$.

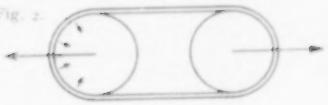
Since the area of the two gaps is $2s^2$, this corresponds to a force of $H^2/(8\pi)$ per cm² of gap area.

This is a very interesting result, which might have been foreseen. If the lateral pressure in the field, whether in air or iron, is equal to $H^2/(8\pi)$, then there is no change in this pressure on passing from the iron into the air at the surface, but as one passes through the winding, H falls to zero and with it the lateral pressure which is taken up by the winding. Hence, on this assumption, the lateral pressure in the iron is transmitted to the winding and there is no separate radial force on the iron itself.

This agrees with the assumption that there is a tension of $H^2/(8\pi)$ along the lines of force and a lateral pressure of $H^2/(8\pi)$, but it does not agree with Maxwell's assumption that the tension along the lines is equal to $BH/(4\pi) - H^2/(8\pi)$. It is difficult to see how Maxwell's assumption can explain the facts. If a rubber band were passed half round a cylinder and stretched as shown in Fig 2 it would exert radial forces on the cylinder, and the resultant of all the radial forces acting on the cylinder must be equal to the sum of the tensions on the two sides. In the magnetic case we find that the resultant of all the radial forces is equivalent to $H^2/(8\pi)$ that is $B^2/(8\pi\mu^2)$ on each square centimetre of the air-gaps and this must surely be equal to the tension in the field. If this is so, then the total force per cm² in the gap $B^2/(8\pi)$ is simply divided into two components, one $B^2(1 - 1/\mu^2)/8\pi$ exerting its force on the pole-face—the force then being transmitted mechanically throughout the iron—and the other $B^2/(8\pi\mu^2)$ which is transmitted magnetically through the iron to exert by catapult action a lateral pressure which is exerted on the windings. This agrees with the result obtained on Moullin's assumption.

Fig. 2 indicates clearly what we mean by the term 'catapult action.'

Fig. 2.



The Correct Solution

Having thus explained the divergent views and the resulting confusion, we shall now show that the matter can be settled quite simply and definitely.

If the air-cored toroid be immersed in a medium of permeability μ, the flux density B and the forces on the conductor, with the same current as before, will be increased to μ times their previous values. Hence the tension in the field and the lateral pressure will also be increased in the same ratio; both will now be $HB/(8\pi)$ or $\mu H^2/8\pi$ per cm².

This is easily confirmed by considering the force between two parallel conductors in a medium of permeability μ. If the two conductors carry equal currents I in opposite directions, as in Fig. 3 (a), the force on each conductor per cm of length due to the field of the other is

$I^2\mu/(100a)$. The value of H at the point P is $4Ia/(10r^2) = 4I\cos^2\theta/(10a)$ and on integrating $\mu H^2/(8\pi)$ over the whole mid-plane, that is from $\theta = -\pi/2$ to $\theta = +\pi/2$ one obtains $I^2\mu/(100a)$ per cm of length of conductor, which agrees with the repulsive force between the conductors.

Similarly, if the currents are in the same direction, as in Fig 3 (b) the value of H at the point P is $4Ib/(10r^2) = 4I \sin\theta\cos\theta/(10a)$, which, on integrating $\mu H^2/8\pi$ from $\theta = -\pi/2$ to $\theta = +\pi/2$ gives $I^2\mu/(100a)$, which again agrees with the force, in this case attractive, between the two conductors. There is, therefore, no doubt whatever that in a medium of permeability μ the longitudinal tension and the lateral pressure are both equal to $\mu H^2/(8\pi)$ or $BH/(8\pi)$ or $B^2/(8\pi\mu)$.

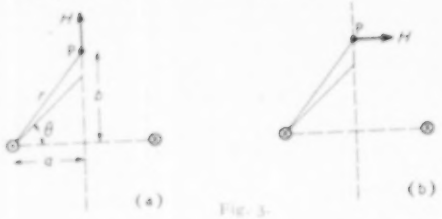


Fig. 3.

As the total force per cm² in the air-gap of the electromagnet is $B^2/(8\pi)$ and the magnetic tension in the iron is $BH/(8\pi)$, the force acting on the pole-face must be the difference, viz.

$$\frac{B^2}{8\pi} - \frac{BH}{8\pi} = \frac{B}{8\pi}(B - H) = \frac{B}{8\pi} \times 4\pi J = \frac{B}{2} \times J$$

Hence the correct result is obtained by multiplying the intensity of magnetization J, not by $(B - H)/2 = 2\pi J$, as Maxwell assumed, nor by $(B + H)/2 = H + 2\pi J$, as Moullin assumed, but by $B/2$.

Turning again to the iron ring of square section with its uniform winding, we can see now that the radial pressure per cm² drops from $BH/(8\pi)$ to $H^2/(8\pi)$ on passing through the surface from the iron into the air, and then from $H^2/(8\pi)$ to zero on passing through the winding.

Of the total force $B^2/(8\pi)$ per cm² in the air-gap, the greater part [viz., $(1 - 1/\mu)B^2/(8\pi)$] acts on the pole-face; the remainder $B^2/(8\pi\mu)$ passes into the iron as a magnetic tension, and causes two catapult forces, one on the iron itself, equal to $(1/\mu - 1/\mu^2)B^2/(8\pi)$, and the other on the winding, equal to $B^2/(8\pi\mu^2)$ per cm².

Much of this is only of academic interest, as $1/\mu^2$ is generally negligible compared with unity, and the magnetic circuit is rarely wound uniformly. If, in the above examples, the winding were limited to the upper half of the ring, even if the total ampere-turns were the same, magnetic leakage would cause the flux in the ring to vary from point to point around the ring.

If the air-gap is negligibly small, the total pull, and the part of it that acts on the pole-face, can be calculated from the flux crossing the gap. If the gap is not negligible, fringing will have to be taken into account. Although there may be no winding on the lower half, the tangential H is the same just outside the ring as it is just inside, and, on passing from iron to air, the lateral stress will fall from $BH/(8\pi)$ to $H^2/(8\pi)$, but as there is no current-carrying winding to take up the latter stress, it will be taken up by the leakage field. We do not propose, however, to go more fully into this aspect of the problem.

G. W. O. H.

ALIGNED-GRID VALVES

Distribution of Current Density

By D. C. Rogers, A.M.I.E.E.

(Standard Telephones and Cables, Ltd.)

SUMMARY.—The electron beam from the cathode of an amplifier valve is focused into well-defined beams by the action of the electric field in the region of the control grid. In aligned-grid valves, use is made of this phenomenon by placing the screen-grid wires in positions where they will intercept only a small fraction of the total cathode current.

Owing to the doubtful validity of the existing theoretical treatment of the subject, an experimental study of the current-density distribution in these electron beams was undertaken for the purpose of determining the best position of the screen wires.

It is often assumed that the grid-to-screen distance should be such that the focus of the electron beam lies in the plane of the screen. The measurements described here show that this is not necessarily true and that a lower screen current will usually be obtained with a smaller grid-to-screen distance.

LIST OF SYMBOLS

a = Grid pitch, measured between centres of grid wires.
d = Grid-wire diameter.
l_g = Cathode-grid distance.
l_a = Cathode-anode distance in a triode, or cathode-screen distance in a tetrode.
$l_a - l_g$ = Grid-anode distance in a triode, or grid-screen distance in a tetrode.
V_g = Grid potential with respect to cathode (e.s.u.)
V_a = Anode potential of a triode or screen voltage of a tetrode with respect to cathode. (e.s.u.)
V_x = Potential in the grid plane. (e.s.u.)
e = Electron charge (e.s.u.)
m = Electron mass.
I = Current flowing between two grid wires, per unit length of grid wire. (e.s.u.)
J = Cathode-current density (e.s.u.)
y = Distance co-ordinate measured parallel to the cathode surface and perpendicular to the grid wires.
z = Distance co-ordinate measured perpendicular to the cathode surface.
V_d = The anode potential of an equivalent diode (e.s.u.) (See Appendix I.)
s = The cathode-anode distance of an equivalent diode. (See Appendix I).
v = Electron velocity (cm/sec)
t = Time in seconds.
$\mathcal{E}$ = Electrostatic field (e.s.u.)
μ = Amplification factor.

1. Introduction

In pentode or beam tetrode valves, it is often found that the power-handling capacity is limited by the maximum permissible screen-grid dissipation and, accordingly, much attention has been devoted to the positioning of the screen wires in the shadow of those of the control grid, in order to intercept a minimum fraction of the cathode current. In receiving-type amplifying valves, too, the reduction of screen current is of considerable importance since the noise current generated by such valves is largely dependent on the amount of current falling on the screen.

It has long been known that in the useful operating range of most valves the electrostatic field in the region of the control-grid causes the electrons to be deflected towards a plane midway between the grid wires. Some typical electron paths in the cathode-grid and grid-screen regions are shown in Fig. 1. Bull[1] has shown that the apertures between the grid wires form cylindrical electron lenses and has shown how their focal lengths can be determined, using the method of Davisson and Calbick[2,3].

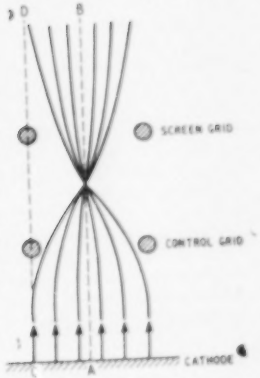


Fig. 1. *Typical electron paths in an aligned-grid valve.*

In practice it is found that the screen current is not reduced to zero, as would be expected if the electron beams came to perfect foci; this fact could be attributed to one or more of the following causes:

(a) The lenses formed by the grid apertures possess large aberrations.
(b) The mutual repulsion between electrons, due to their charge, causes a tendency for the beam to diverge in the grid-screen region.
(c) The electrons are emitted from the cathode with finite velocities, distributed according to the Maxwell law.

MS accepted by the Editor, June 1949

(d) The electrostatic field configuration in the vicinity of the screen grid is such as to deflect the electrons towards the screen wires.

In Appendix I it is shown that the space-charge effect mentioned in (b) above is insufficient to account for observed screen currents, and it can readily be demonstrated (*for* example, by observing the variation in the proportion of cathode current intercepted by the screen as the cathode temperature is varied) that electron emission velocity is also not a major cause of the spread of the beams. It must therefore be concluded that the screen currents obtained in practice result from the existence of large aberrations in the lenses, from the effect of the electrostatic field in the screen region, or from both these causes.

Now the method employed by Bull to determine the focal length is, as he points out, valid only for electrons travelling close to the central plane AB (Fig. 1). It is clear, however, that the screen current does not arise from these electrons, but from those passing very near to the control-grid wires and we are, therefore, justified in asking whether the focal plane is, in fact, the optimum position for the screen. In view of the difficulty of obtaining an analytical solution for the trajectories of electrons passing near the control-grid wires, involving as it would

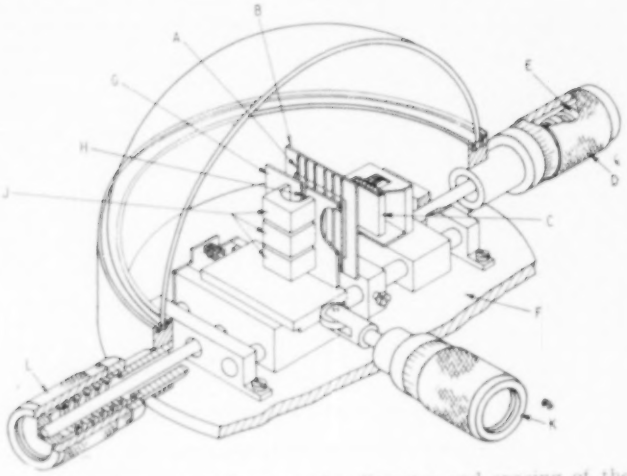


Fig. 2. Simplified sketch of experimental valve.

consideration of the effect of space charge on the field in the grid region, it was decided that an experimental approach would be profitable. The investigation to be described here was accordingly undertaken; viz., an investigation of the distribution of current density in the plane of the screen.

This current density distribution depends on the trajectories of individual electrons in their flight from the cathode to the screen plane, and hence on the field configuration in the grid region. It does not depend appreciably on the geometry of the screen, except in the immediate proximity of a screen wire, since the field irregularities do not extend far into the grid-screen space. In making these measurements, details of which are given in a later section, the screen was replaced by a solid metal plate provided with a narrow slot, the current passing through the slot thus being a measure of the current density falling on the plate. The measured distribution is, therefore, that appertaining to a particular cathode-grid geometry and grid-screen distance, and does not include the localized effect around the screen wires. The measurements were carried out for a range of grid-screen distances and with various cathode-grid geometries.

It should be noticed that while this method enables the optimum screen position to be determined, it does not enable the screen current at this position to be predicted. The screen current will, in general, depend on the conditions obtaining in the screen-anode region (e.g., the potential and location of the anode and the space-charge conditions in the intervening space) as well as on the diameter and spacing of the screen wires.

2. Experimental Procedure

The measurements were made on a continuously-evacuated demountable valve in which the electrodes could be moved by manipulation of external micrometer controls while the valve was in operation. In the interests of simplicity attention has been confined to parallel plane electrode structures. However, little error should arise in the application of the results to practical valve designs, since the grid-support rods effectively divide the valve into two sections, each of which approximates roughly to a parallel plane structure.

A simplified sketch of the apparatus is shown in Fig. 2. The grid consisted of straight molyb-

denum wires A welded to a rectangular steel frame B. The cathode C was of the indirectly-heated oxide-coated nickel type, the coated area being a flat face 13 mm × 10 mm. The cathode-grid distance could be adjusted by means of the micrometer screw D. The vacuum was maintained by the flexible metal bellows E soldered to the metal envelope F of the valve.

As mentioned in the introduction, the screen was replaced by a flat metal plate, shown at G in the figure. This plate was provided with a slot H, 0.1-mm wide, parallel to the grid wires, behind which was located a collector electrode J, operated at a positive potential of 150 volts with respect to the anode G. The current flowing to this collector electrode was a measure of the current density falling on the anode plate. The whole plate could be moved in a direction parallel to the grid plane and perpendicular to the grid wires by the micrometer screw K. Thus by observing the variation of the collector current as the slot in the anode was moved from a position behind one grid wire to behind the next, the variation in current density across the electron beam could be determined.

The anode-grid spacing was also variable and was controlled by micrometer L. In order to avoid spurious effects at the ends of the slot, the collector was divided into three sections as shown, and measurements were made of the current to the centre section only, the outer two sections being maintained at the same potential as the centre section. Since under some conditions the current reaching the collector amounted to a few microamperes only out of 2 total cathode current of 4 mA it was necessary to prevent electrons from reaching the collector by other than the intended paths ; for instance, by secondary emission from various parts of the apparatus. This was accomplished by suitably placed shields at cathode potential : these have been omitted from Fig. 2 for the sake of clarity.

The experimental procedure was to scan the beam emerging from the centre grid aperture by moving the collector slot transversely and measuring the collector current at intervals of 0.1 mm. This process was repeated for various anode-grid distances, the total anode current being maintained constant at 4 mA by adjusting the anode voltage. Because of the large number of variables that would otherwise be involved, measurements have been made with the grid at cathode potential only. A curve showing the ratio of the collector current at each point in the beam section to the average across the beam was plotted for each anode-grid distance. This ratio is equal to the ratio of the current density at that point to the average current density. A typical series of such curves is shown in Fig. 3.

It is readily shown that, provided the effects of initial emission velocity are negligible, the shapes of the electron trajectories are dependent only on the shape of the electrodes and not on the actual dimensions. Hence, any geometrical scaling factor can be used without changing the relative distribution of current density. It is, therefore, convenient to maintain one dimension constant and to relate the other dimensions to this fixed dimension. In this case the grid pitch a was kept constant at 2 mm and other dimensions were in terms of grid pitch. The experiments were repeated for various values of the ratio l_a/a (cathode-grid distance/grid pitch) and d/a (grid-wire diameter/grid pitch) likely to be encountered in practice. (See also list of symbols).

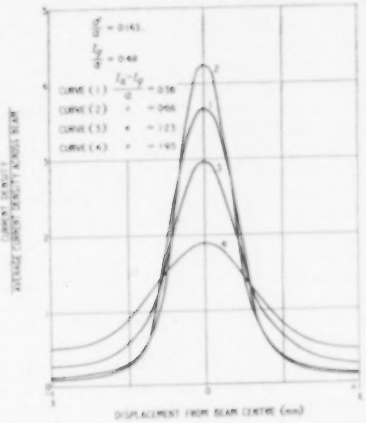

$\frac{d}{a}$ = 0.145
$\frac{l_g}{a}$ = 0.49
CURVE (1) $\frac{l_a-l_g}{a}$ = 0.38
CURVE (2) " = 0.66
CURVE (3) " = 1.23
CURVE (4) " = 1.97
RATIO CURRENT DENSITY/MEAN VALUE
DISPLACEMENT FROM BEAM CENTRE (mm)

Fig. 3. *Variation in current density across the electron beam.*

The shape of the electron paths is dependent also on the ratio only of the electrode potentials and not on their magnitude. However, since the grid was maintained at cathode potential, this amounts to saying that the relative distribution of current density is independent of anode voltage.

3. Results and Their Significance

Referring again to Fig. 3, which shows a typical series of curves of current distribution in the anode plane (or in what would be the screen plane, were the valve a tetrode) it will be seen that, as would be expected, the current density rises from a minimum behind the grid wires to a maximum value midway between them.

These curves were obtained with $d/a = 0.145$ and $l_g/a = 0.49$. As the ratio grid-anode distance/grid pitch, $(l_a - l_g)/a$ is increased, the magnitude

of the maximum between the grid wires rises until it is itself a maximum and then falls again, the maximum in this case occurring when $(l_a - l_g)/a = 0.65$. This is made clearer by reference to Fig. 4, in which the dotted curve shows the variation with $(l_a - l_g)/a$ in current density in the mid-plane between the grid wires; i.e., along the line **AB**, Fig. 1.

Quite obviously the value of $(l_a - l_g)/a = 0.65$, which gives a maximum in this curve, corresponds to the case when the focus of the electron beam is in the plane of the anode or, of course, the screen, had the valve been a tetrode. It is interesting to note that, with the electrode geometry corresponding to this point, the position of the focus as determined by Bull's method would have been at a distance $0.45a$ from the grid plane.

Now let us examine the variation in current density behind the grid wires, and in the plane of the anode of the experimental triode (i.e., along the line CD, Fig. 1) as the anode-grid distance is varied. This current is given by the points X on the curves of Fig. 3, but is shown more clearly by the solid curve of Fig. 4, which has been plotted from Fig. 3. It will be seen that there is no minimum corresponding to the maximum occurring along AB, but rather that the current density increases progressively as the distance between grid and anode is increased.

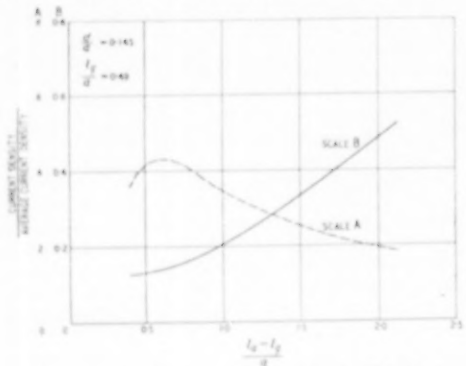


Fig. 4. *Variation of current density along the line AB of Fig. 1 (dotted curve) and along the line CD (solid curve).*

The implication of this is clearly that for minimum screen current, at least at zero grid voltage, the screen should be located as nearly as possible to the control grid, and not in the plane of the focus, as is frequently assumed. However, before considering this point further, the effect of varying the other valve dimensions will be described.

Fig. 5 shows a family of curves similar to those of Fig. 4, plotted for various values of the grid-wire diameter/grid pitch parameter d/a. It will be seen that the value of $(l_a - l_g)/a$ for which the beam focuses in the anode plane decreases as the grid-wire diameter is increased. This

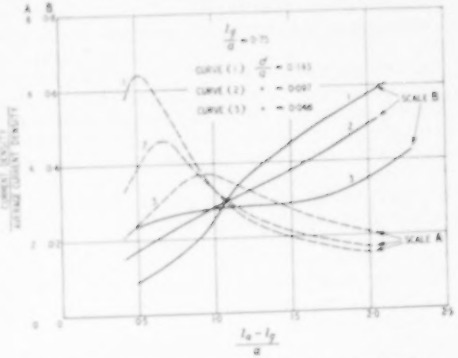


Fig. 5. *Variation of current density along the line AB (dotted curves) and along the line CD (solid curves) for various values of d/a.*

is also shown in Fig. 6, which shows the value of $(l_a - l_g)/a$ for which the beam focuses in the anode plane versus the grid-wire diameter parameter d/a.

The effect on the current density behind the grid wires is, as shown in Fig. 5, that for small values of $(l_a - l_g)/a$ the current density decreases with increasing wire diameter, while for large values of $(l_a - l_g)/a$ the reverse is the case.

Fig. 7 shows another series of curves, also similar to Fig. 4, plotted for a constant grid-wire diameter, corresponding to $d/a = 0.048$ and for varying cathode-grid distances. It will be seen that the value of $(l_a - l_g)/a$ for focus in the anode plane is substantially independent of cathode-grid distance, but that the value of the current density maximum increases as the cathode-grid distance increases. It should be mentioned here that these maxima of current density may be quite appreciably in error, since the measuring slot has finite width, and this width may be greater than the width of the high-current-density part of the beam in cases where the focus is very sharp.

In Appendix II an expression is developed for the focal length of the lens formed by the grid apertures, using a similar method to that used by Bull, but based on electrostatic considerations only. The simplicity of the expression obtained in this way draws attention to the fact that the dominant feature in determining the focal length at zero grid voltage is the ratio of

grid-wire diameter to grid pitch, a fact which is strikingly confirmed by the measurements.

The expression obtained is

$$f = \frac{3}{2} f_o \left[1 + \frac{1}{6} \frac{f_o}{l_g} \right]$$

$$\text{where } f_o = \frac{a}{\pi} \log \frac{a}{\pi d} \quad .. \quad .. \quad .. \quad (1)$$

In Fig. 6 the calculated value of f is plotted as a function of d/a for $l_g/a = 0.75$, the experimental results being shown for comparison. For reasons outlined in the appendix, the apparently good agreement between the calculated and measured values is probably due to a fortuitous cancellation of errors.

The effect of change in cathode-grid distance on the current density behind the grid wires is also shown in Fig. 7. At small values of $(l_a - l_g)/a$ the current density decreases with increase of l_g/a, while with larger values of $(l_a - l_g)/a$, it increases with l_g/a.

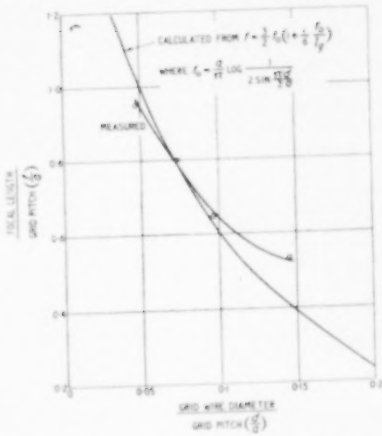


Fig. 6. *Value of $(l_a - l_g)/a$ for which the beam focuses in the anode plane; both curves relate to $l_g/a = 0.75$.*

An interesting series of curves is shown in Fig. 8. These are similar to Fig. 3, and show the current-density distribution across the electron-beam section for thin grid wires. When the anode is very close to the grid [i.e., for $(l_a - l_g)/a$ small] the distribution of current is seen to take the shape of a double-humped curve. This provides additional confirmation, if any is necessary, that the lenses formed by the grid apertures have large aberrations, for it is clear that in this case the outer electrons of the beam are subject to a stronger lens action than those in the centre, resulting in a 'crowding' of electrons at the outer edges of the beam.

Let us now consider whether these results can be applied to the design of a practical valve with reduced screen current. The results apply only to zero grid voltage, of course, whereas in

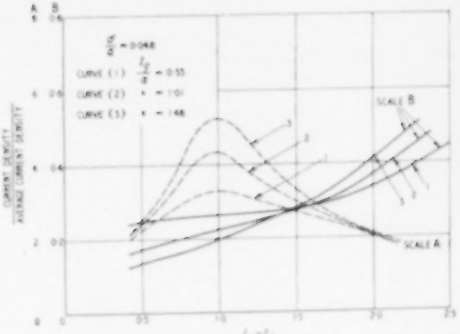


Fig. 7. *Variation of current density along the line AB of Fig. 1 (dotted curve) and along the line CD (solid curves) for various values of l_g/a.*

use the valve will normally have an alternating voltage applied to the grid, and it is the average screen current over the grid-voltage cycle which must be minimized. However, if we confine our attention to amplifiers in which the grid is not driven positive, the conditions for minimum average screen dissipation and minimum dissipation at the instantaneous current maximum would not be expected to be very different.

It is clear from Figs. 3 and 4 that to intercept the minimum cathode current, the screen-grid wires should be as near to the control grid as possible. Reducing the control-grid-to-screen distance will also have the effect of reducing the grid-to-screen amplification factor of the valve, and since, in general, it will be necessary for the static characteristics of the valve to be within specified limits it will be necessary to correct this reduction in amplification factor by increasing the diameter of the grid wires. As shown in Fig. 5, if the grid-screen distance is less than the grid pitch, the effect of large diameter grid wires will be to reduce the screen current still further. Fig. 7 shows that, with large grid wires and small grid-screen distances, the screen current will not vary rapidly with cathode-grid distance, and hence this distance need depend only on the static characteristic requirements.

The limit to the reduction in control-grid-to-screen distance is set by practical considerations. Also, when the cathode-grid distance is small, a

small grid-to-screen distance and large diameter grid wire result in non-uniformity of the field at the cathode and consequent increased curvature of the grid-volts–anode-current characteristics.

It will be appreciated that even if these findings were applicable to positive grid voltages, they would not be of much value. A grid of large diameter wires operated at a positive voltage would itself intercept an intolerably high proportion of the cathode current.

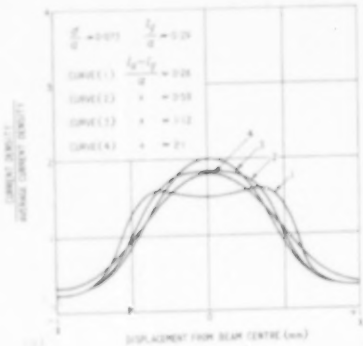


Fig. 8. *Variation of current density across beam section for thin grid wires and small value of I_a/a.*

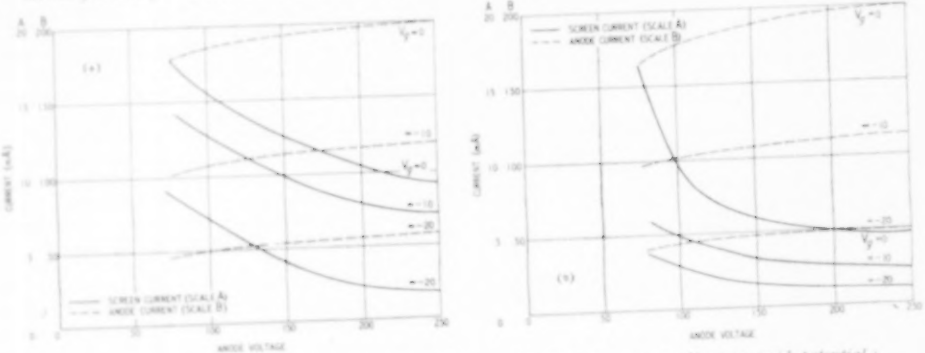


Fig. 9. *Screen and anode currents of (a) valve similar to the Type 807 at 250-V screen-grid potential; (b) valve with thicker grid wires and decreased grid-to-screen distance; screen-grid potential = 200 V.*

Next let us consider whether the results of these experiments can be used to predict the screen current when the valve geometry is known. It is convenient to regard the screen current as consisting of two components, a forward component arriving from the cathode-grid region, and a reverse component consisting of electrons returned from the screen-anode region. The first-mentioned component will be determined by the current density flowing in the region in which the screen wire is located, and by the shape of the electrostatic field in the immediate vicinity of the screen wire. Spangenberg[4], in a theoretical investigation of the positive-grid triode, has shown that in the case of equal grid and anode voltages the positive grid may be considered to intercept current over an area that normally lies between 1.25 and 1.8 of the projected area of the wire. It is possible that Spangenberg's method could be adapted to give an approximate solution for the tetrode, by considering the tetrode control-grid plane as the cathode of Spangenberg's triode, but since the method is based on electrostatic considerations only, it could not be used when the space-charge in the screen-anode region is appreciable. It could be applied, therefore, only to the case of small screen-to-anode distance and high anode voltage, and would not be valid in the more interesting case of the tetrode in which space-charge in the screen-anode region is used to suppress secondary emission from the anode. Furthermore, the process of treating the control grid as the electron source would be valid only for large grid-to-screen distances, whereas it has been proposed above that the method of reducing screen current is to reduce this distance.

The other component of screen current (viz., that returned from the screen-anode region) will depend only on conditions existing beyond the screen. In most beam tetrodes this component will predominate only at very low anode voltages, while the first-mentioned component will predominate at high anode voltages. Any improvement effected by reducing the control-grid-to-screen distance will, therefore, be apparent mainly at high anode voltages.

In order to verify the general conclusions

reached, the following experiment was carried out. A valve was made similar to the type 807 beam power tetrode, but in which the ratio screen-grid distance/grid pitch was reduced from 1.49 to 0.58 and the grid-wire diameter was increased such that d/a was 0.192 instead of 0.096 for the normal valve. It was hoped in this way to make a valve having similar characteristics to the 807 in the negative-grid region, but with reduced screen current. In the modified valve, the screen current was 5.0 mA at $V_a = 250$ V, $V_{g1} = 0$, $I_a = 200$ mA. (I_a set to 200 mA by adjusting the screen voltage), compared with 10 mA for the normal valve. Owing to the fact that the increase in grid-wire diameter did not quite compensate for the decrease in grid-screen distance, the anode current of 200 mA was obtained at a somewhat lower screen voltage than normal; viz., 200 V instead of 250 V. This resulted in the space-charge conditions in the screen-anode space being altered with a consequent slight change in the shape of the anode-volts–anode-current curve. The characteristic curves of the modified and the normal valve are shown in Figs. 9(a) and (b) respectively for comparison. In the modified valve, some increase in the curvature of the grid-voltage–anode-current characteristic was noticeable.

4. Conclusion

These experiments have shown that some reduction in the screen current in an aligned-grid valve should be possible by the use of a smaller grid-to-screen distance than is common at present, provided this reduced spacing is not incompatible with conditions set by other characteristics of the valve. This conclusion is in accord with physical reasoning, for it is clear that there must be regions of very low current density immediately behind the control-grid wires, since the only electrons able to reach these regions—apart, of course, from those returned from the screen-anode space at low anode voltages—are those suffering deflections of nearly 90° at the control grid.

5. Acknowledgment

The author wishes to thank Mr. M. T. G. Chick for valuable assistance in the experimental work, and Messrs. Standard Telephones & Cables Limited for permission to publish this paper.

APPENDIX I

Space-charge defocusing.

This appendix is intended to show that defocusing of the electron beams, due to space charge, does not cause appreciable spreading of the beam. It is convenient to consider the case of an aberration-free lens of such focal length that the beam would form a line focus in the plane of the screen in the absence of space-charge. The presence of space-charge will cause the beam to have finite thickness in the screen plane, and it is our aim to find this beam thickness.

An electron will emerge from the grid plane with a velocity v_s given by

$$v_s = \sqrt{\frac{2eV_s}{m}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (2)$$

where V_s is the effective potential in the grid plane in electrostatic units.

The electron then travels under the influence of an accelerating field towards the anode. This field will be assumed uniform and equal to

$$\mathcal{E} = \frac{V_a - V_s}{l_a - l_g} \text{ electrostatic units} \quad \ldots \quad \ldots \quad (3)$$

The distance travelled in time t is

$$\sqrt{\frac{2eV_s}{m}}\, t + \tfrac{1}{2}\frac{e}{m}\frac{(V_a - V_s)}{(l_a - l_g)}t^2 \text{ cm} \quad \ldots \quad \ldots \quad (4)$$

so that the time t_a taken to reach the screen plane is given by

$$(l_a - l_g) = \sqrt{\frac{2eV_s}{m}}\, t + \tfrac{1}{2}\frac{e}{m}\frac{(V_a - V_s)}{(l_a - l_g)}t^2 \quad \ldots \quad (5)$$

whence

$$t_a = \sqrt{\frac{2m}{e}}\frac{(\sqrt{V_a} - \sqrt{V_s})(l_a - l_g)}{(V_a - V_s)} \quad \ldots \quad \ldots \quad (6)$$

Now the 'half-width' y_t of the beam at time t is given by Thompson and Headrich.[5] Expressed in the symbols used here it is:

$$y_t = y_s - v_s t + 2\pi I v_s \frac{m}{e}\frac{(l_a - l_g)^2}{(V_a - V_s)^2}\left(1 + \frac{(V_a - V_s)et}{v_s(l_a - l_g)m}\right)$$

$$\left[\log\left(1 + \frac{(V_a - V_s)et}{v_s(l_a - l_g)m}\right) - 1\right] + \frac{2\pi I v_s (l_a - l_g)^2}{(V_a - V_s)^2}\frac{m}{e} \quad (7)$$

where y_s is the initial 'half-width' of the beam,

I is the current in the beam per cm of beam breadth, i.e., measured parallel to the cathode and to the grid wires,

v_s is the initial velocity of the outermost electrons in a direction perpendicular to the grid wires and to the cathode face.

If the focal length of the lens is such that, in the absence of space-charge, a line focus would be formed in the screen plane,

$$y_s = v_s t_a \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

Now substituting in (7) for v_s, t_a, and v_s from (2), (6) and (8) and rearranging, we get

$$y_{ta} = \frac{2\pi I\sqrt{\frac{2mV_s}{e}}(l_a - l_g)^2}{(V_a - V_s)^2}\left\{\frac{\sqrt{V_a}}{\sqrt{V_s}}\left[\log\frac{V_a}{V_s} - 1\right] + 1\right\} \quad (9)$$

Hence the ratio total beam width/grid pitch

$$= 2y_m/a$$

$$= \frac{4\pi J}{\sqrt{\frac{2e}{m}}} \frac{\sqrt{V_a}(I_a - I_g)^2}{(V_a - V_g)^2} \left\{ \frac{\sqrt{V_g}}{\sqrt{V_a}} \left[\log \frac{\sqrt{V_g}}{\sqrt{V_a}} - 1 \right] + 1 \right\} \quad (10)$$

where J is the average current density over the cathode surface ;

$$\text{i.e., } J = \frac{I}{a} \quad (11)$$

It is well known that

$$J = \frac{\sqrt{2}}{9\pi} \sqrt{\frac{e}{m}} \frac{V_D^{3/2}}{S^2} \quad (12)$$

where V_D is the effective anode potential of the so-called equivalent diode and S is the effective cathode-anode distance of this diode.

The values of V_D and S have been the subject of much discussion. It will be sufficiently accurate here to take

$$V_D = V_g + V_a/\mu \quad (13)$$

and

$$S = l_g + l_a/\mu \quad (14)$$

Substituting for J from (12), (13) and (14) and considering the case when $V_g = 0$, we obtain

$$2y/a = \frac{8}{9\mu^2} \frac{(l_a - l_g)^2}{(1 - V_g/V_a)^2 (l_g + l_a/\mu)^2} \left[\left(\log \sqrt{\frac{V_g}{V_a}} \right) + \sqrt{\frac{V_g}{V_a}} - 1 \right] \quad (15)$$

$\frac{V_g}{V_a}$ can be obtained by Bull's method [1] or, alternatively, from electrostatic considerations only, from Appendix II, equation (17).

For example, when $l_a = 3l_g$

$$a = l_g$$

$$d/a = 0.1$$

then $\mu = 10.6$

and $V_g/V_a = 0.0735$

so that $2y/a = 0.042$.

It is thus clear that the beam spread, due to space-charge, is not sufficient to account for the values of screen current observed in practice.

APPENDIX II

Davisson & Calbick's equation for the focal length of a cylindrical electrostatic lens is

$$\frac{1}{f_a} = \frac{\mathcal{E}_2 - \mathcal{E}_1}{2V_a} \quad (16)$$

where V_a is the potential in the aperture, and $\mathcal{E}_1$ and $\mathcal{E}_2$ are the fields on the entrance and exit sides of the lens respectively.

The effective potential in the grid-plane of a parallel plane triode at zero grid voltage is given by

$$V_a = V_a \frac{I_g}{I_a + \mu I_g} \quad (17)$$

provided the interelectrode spacings are not small compared with the grid pitch. The relevant values of $\mathcal{E}_1$ and $\mathcal{E}_2$ are those at the cathode and anode, respectively.

$$\text{Hence } \mathcal{E}_1 = \frac{V_g}{I_g} = \frac{V_a}{I_a + \mu I_g} \quad (18)$$

$$\text{and } \mathcal{E}_2 = \frac{V_a - V_g}{I_a - I_g} = \frac{V_a}{I_a - I_g} \left(1 - \frac{I_g}{I_a + \mu I_g} \right) \quad (19)$$

From (16), (17), (18) and (19), we obtain after simplification

$$f_a = \frac{2(I_a - I_g)}{\mu} \quad (20)$$

For moderately thin grid wires,

$$\mu = \frac{2\pi (I_a - I_g)}{a \log \frac{1}{2 \sin \frac{\pi d}{2a}}} \quad (21)$$

$$\text{whence } f_a = \frac{a}{\pi} \log \frac{1}{2 \sin \frac{\pi d}{2a}} \quad (22)$$

In this expression f_a is the focal length of the lens formed by a grid aperture. This focal length is not the distance between the grid plane and the line focus of the beam, however, since, after leaving the lens region, the electrons travel along parabolic paths under the influence of a uniform accelerating field.

The distance f from the lens to the point where the beam focuses is shown by Bull to be

$$f = f_a \left[1 + \tfrac{1}{4} \frac{f_a}{(I_a - I_g)} \left(\frac{V_a}{V_a} - 1 \right) \right] \quad (23)$$

$$= f_a \left[1 + \tfrac{1}{4} \frac{f_a}{(I_a - I_g)} \left(\frac{I_a + \mu I_g}{I_a} - 1 \right) \right] \quad (24)$$

$$= \tfrac{5}{4} f_a \left[1 + \tfrac{1}{5} \frac{f_a}{I_a} \right] \quad (25)$$

f/a has been plotted versus d/a for $I_a/a = 0.75$ in Fig. 6, and it will be seen that there is reasonably good agreement with the measured figures. A possible explanation of this unexpected agreement is as follows :—

In calculating f, V_a was taken to be the average potential in the grid plane, whereas the potential mid-way between the grid wires should have been used ; this, in the absence of space-charge, would be greater than the average potential. On the other hand, the effect of space-charge would be to reduce the potential between the grid wires, and thus the errors arising from the two sources are seen to be in opposite senses.

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TRIPLET REFLECTOR ARRAY

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SUMMARY.—The paper describes a reflector system using parasitic elements arranged in groups of three spaced $\lambda/8$ apart. A theoretical discussion is given, and it is shown that the polar diagrams calculated theoretically are realized with a high degree of accuracy in practice.

Most of the experimental work has been carried out at a wavelength of approximately 6 metres, though the theory is equally applicable to other wavelengths.

In the paper comparisons are drawn between the proposed triplet-reflector arrays and the corresponding arrays having a single reflector element behind each driven element. The relation between reflector reactance required by theory and reflector length found in practice is discussed.

1. Introduction

DURING the development of early-warning radar in the early years of the war, the author took part in the setting up of high-power transmitting aerials. The present suggested design of reflector system arose from a consideration of the difficult problem of minimizing the backward radiation from the transmitting-aerial systems used on the so-called CH radar stations. The object of the design was to obtain a reflector system less sensitive to frequency changes than those already in use and, therefore, demanding less strict tolerances in the cutting of reflector wires to length. Further, it was hoped to secure a better back-to-front ratio than was obtainable using arrays with single reflector elements. It was found, in fact, that no improvement over an array with single reflectors in perfect adjustment (a very critical state of affairs to achieve) was obtained, but that the adjustment of the triplet array was less critical and that the optimum conditions could be realized over a wider band of frequencies.

2. Outline of C.H. and Triplet Arrays

The CH-aerial system was constructed of units consisting of a horizontal half-wave dipole and parasitic reflector spaced 0.17λ apart. The length

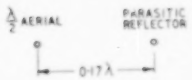

Fig. 1. *Horizontal elements of CH system viewed end on.*

of the reflector was very critical and of the order of $\lambda/2$ (Fig. 1). Such units were stacked vertically at half-wave intervals to form arrays, the number of units in a stack being decided by the wavelength and the height of tower available. In practice this number varied between 4 and 12.

It was suggested that an improvement in performance could be obtained by having three identical reflectors, spaced $\lambda/8$ from one another vertically, behind each 'excited' element (Fig. 2). This is the basis of the triplet system, and these units can be stacked in the same way to form arrays.

MS accepted by the Editor, April 1949

3. Theory of Single Stack Systems

A detailed theoretical treatment of single- and two-stack arrays has recently been given by Walkinshaw.[1] It is only necessary, therefore, to indicate briefly how the polar diagrams of such aerials are calculated in order that the method of calculation might be compared with that used for the triplet array.

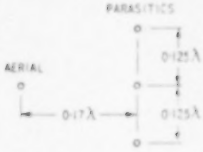

Fig. 2. *End view of triplet reflector.*

3.1 Let us consider first the case of the single $\lambda/2$ dipole and reflector in free space (Fig. 3) where :—

I_0 = current in dipole
I_1 = current in reflector
Z_m = mutual impedance between dipole and reflector
d = distance between dipole and reflector
Z = self-impedance of reflector.

If the e.m.f. induced in the reflector $= I_0 Z_m$

$$I_1 = \frac{-I_0 Z_m}{Z} \text{ and } I_1 Z + I_0 Z_m = 0$$

hence

$$\frac{I_1}{I_0} = -\frac{Z_m}{Z} = A \angle \alpha \text{ say, where } \alpha = \text{angle of lead of } I_1 \text{ on } I_0$$

To calculate the shape of the polar diagram at any distant point P in the plane of the array [Fig. 3(b)], it is convenient to take the aerial current as unity with zero phase angle, and reflector current as $A \angle \alpha$

At P the lead of the reflector current on the aerial current is $\alpha - (2\pi d/\lambda)\cos\theta = \beta$, say.

Then the field at P is proportional to

$$\cos\theta(1 + A\angle\beta)$$
$$= \cos\theta(1 + A\cos\beta + jA\sin\beta)$$

And the amplitude of the field strength is given by

$$\cos\theta\sqrt{1 + 2A\cos\beta + A^2}$$

The term $\cos\theta$ is used as an approximation to the polar pattern of a half-wave dipole $\frac{\cos(\pi/2\sin\theta)}{\cos\theta}$ and is quite in keeping with the order of accuracy that can be realized from any experimental polar diagrams of backward radiation.

For thin wire half-wave elements Z is taken as $80 + jX$ ohms. A family of polar diagrams can be plotted by altering the value of X; i.e., by varying the length of the reflector. It is found that there is negligible change in the shape of the lobe of forward radiation as X is altered but that the backward radiation changes considerably both in value and lobe shape.

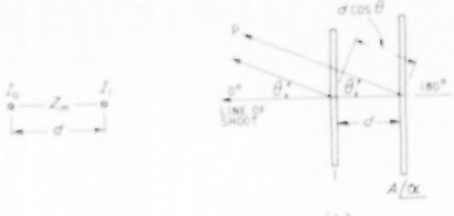

Fig. 3. *Elevation (a) and plan (b) of single dipole and reflector.*

Three polar diagrams have been plotted (see Fig. 4) for X having the values + 40, + 60, + 80 ohms. From the graph of back-to-front ratio versus reflector reactance (Fig. 5), the condition for minimum radiation on the bearing 180° can be seen to occur at $X = 60$ ohms.

3.2. The single-stack triplet-reflector system is treated in the same way as the above.

Let I_0 = dipole current

I_1, I_2 = currents in reflectors

Z_{m1}, Z_{m2}, Z_{m3}, Z_{m4} = mutual impedances between the various elements as shown in Fig. 6.

Z = self-impedance of any of the reflectors (all identical).

By equating the e.m.fs induced in the reflectors

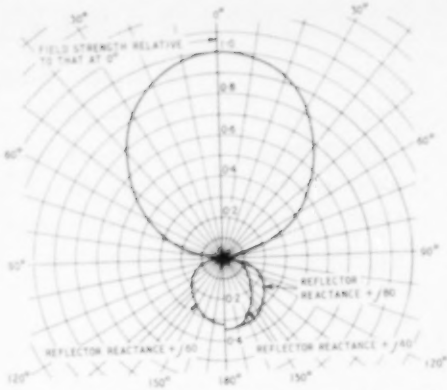

Fig. 4. *Theoretical polar diagram of single-stack array with single reflector.*

to the products of their respective currents and self-impedances, we find that

$$0 = I_0 Z_{m1} + I_1(Z + Z_{m4}) + I_2 Z_{m3}$$
$$0 = I_0 Z_{m2} + 2I_1 Z_{m3} + I_2 Z$$

hence

$$\frac{I_1}{I_0} = -\frac{ZZ_{m1} - Z_{m2}Z_{m3}}{Z(Z + Z_{m4}) - 2Z_{m3}^2} = A_1\angle\alpha, \text{ say}$$

and

$$\frac{I_2}{I_0} = -\frac{Z_{m2}(Z + Z_{m4}) - 2Z_{m1}Z_{m3}}{Z(Z + Z_{m4}) - 2Z_{m3}^2} = A_2\angle\alpha_2, \text{ say.}$$

Putting $\beta_1 = \alpha_1 - (2\pi d/\lambda)\cos\theta$ and $\beta_2 = \alpha_2 - (2\pi d/\lambda)\cos\theta$

the shape of the horizontal polar diagram is readily seen to be given by the expression

$$\cos\theta(1 + 2A_1\angle\beta_1 + A_2\angle\beta_2)$$

whose amplitude is

$$\cos\theta\sqrt{(1 + 2A_1\cos\beta_1 + A_2\cos\beta_2)^2 + (2A_1\sin\beta_1 + A_2\sin\beta_2)^2}$$

For ease of numerical evaluation this can be written

$$A_2\cos\theta\sqrt{\left\{\frac{1}{A_2} + \frac{2A_1}{A_2}\cos\beta_1 + \cos\beta_2\right\}^2 + \left\{\frac{2A_1}{A_2}\sin\beta_1 + \sin\beta_2\right\}^2}$$

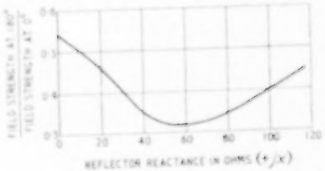

Fig. 5. *Back-to-front ratio of single stack with single reflector.*

Back-to-front ratio versus reflector reactance is plotted in Fig. 7 and three selected polar

diagrams in Fig. 8. As might be expected there is a marked decrease in backward radiation by employing three reflectors instead of one, but more noteworthy is the increase in bandwidth, for nowhere does the backward radiation exceed 0.3 of the maximum forward radiation within the range of reflector reactances + 80 to + 200 ohms. This compares very favourably with the 40-ohm tolerance permissible on the single-stack reflector system to keep the back-to-front ratio down to 0.35.

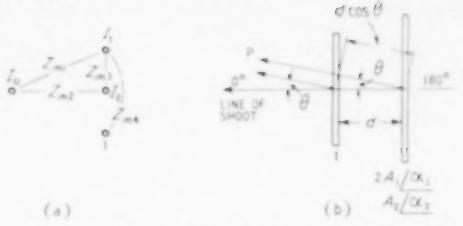

Fig. 6. *Elevation (a) and plan (b) of dipole and triplet reflector.*

4. Two-Stack Arrays

It will be seen from an examination of the polar diagrams calculated by Walkinshaw that the behaviour of two-stack arrays differs in several important respects from that of a single-stack array. The behaviour of arrays with a larger number of elements, however, has been shown to approximate very closely to that of the two stack. Calculations, therefore, have been limited to arrays having two stacks and the comparison which follows is between such arrays having single and triplet reflectors.

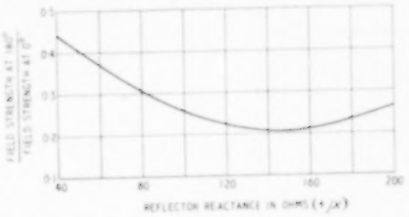

Fig. 7. *Back-to-front ratio of single stack with triplet reflector.*

4.1 *Two-stack Single Reflector Array*

The various currents and mutual impedances are as shown in Fig. 9. Then, as before

$$0 = I_0 Z_{m1} + Z_{m2} + I_1(Z + Z_{m3})$$

hence

$$\frac{I_1}{I_0} = -\frac{Z_{m1} + Z_{m2}}{Z + Z_{m3}} = A \angle \alpha$$

and the field strength at any angle θ (in the horizontal plane) to the line of shoot is proportional to $\cos\theta \sqrt{1 + 2A \cos\beta + A^2}$, where β has the same significance as in Section 3.

Some calculated backward polar diagrams are shown in Fig. 10. The back-to-front ratio deteriorates far more rapidly than for a single-stack array as X deviates from its optimum value (see Fig. 11). For values of X not far off optimum, the ratio is much smaller than can be obtained with a single stack; e.g., a ratio less than 0.12 is maintained over the reactance range + *j*40 to + *j*160 ohms.

Another feature which does not occur with single-stack arrays is the appearance of the so-called 'clover-leaf' pattern, obtained when $X = 40$ ohms. The backward radiation is split into three lobes having equal maxima.

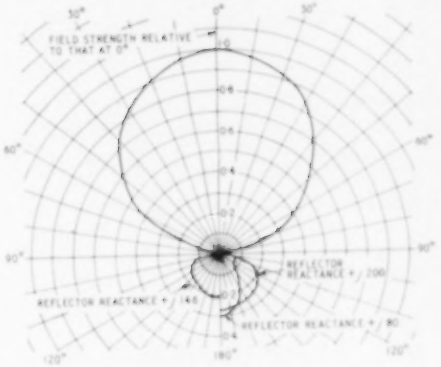

Fig. 8. (Above) *Theoretical polar diagram of single stack with triplet reflector.*

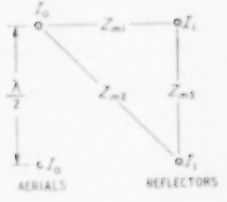

Fig. 9. (Right) *Elevation of two-stack single-reflector array.*

In suppressing backward radiation the field strength at 180° is no longer the only criterion, for once the clover-leaf pattern has been reached any further increase in length (i.e., of inductive reactance) of the reflectors causes the maximum value of the side lobes to rise above the value at 180°.

4.2 *Two-stack Triplet Reflector Array*

The additional reflectors in the triplet system complicate considerably the mathematics involved in calculating the polar diagrams. It is fortunate

that it has been found that multiple-stack arrays approximate very closely as regards polar diagram, to two-stack arrays. Certainly this has only been verified for arrays using single reflectors, but there is no reason to believe that the behaviour would be different for arrays having triplet reflectors. The numerical evaluation of the polar diagram of an array having more than two stacks of triplet reflectors would be extremely complicated and tedious.

The currents, mutual impedances and the vertical spacing are shown in Fig. 12. The other dimensions are as given in Fig. 2.

For reflectors 1, 2 and 3, we find, by equating the induced e.m.fs to the products of the respective reflector currents and self impedances, that :—

$$\begin{aligned} 0 &= I_0(Z_{m2} + Z_{m7}) + I_1(Z + Z_{m11}) \\ &\quad + I_2(Z_{m3} + Z_{m10}) + I_3(Z_{m4} + Z_{m9}) \\ 0 &= I_0(Z_{m1} + Z_{m6}) + I_1(Z_{m3} + Z_{m10}) \\ &\quad + I_2(Z + Z_{m9}) + I_3(Z_{m2} + Z_{m8}) \\ 0 &= I_0(Z_{m2} + Z_{m5}) + I_1(Z_{m4} + Z_{m9}) \\ &\quad + I_2(Z_{m3} + Z_{m8}) + I_3(Z + Z_{m4}) \end{aligned}$$

For brevity, putting

$$\begin{aligned} Z + Z_{m11} &= B \qquad Z + Z_{m9} = C \qquad Z + Z_{m4} = D \\ Z_{m3} + Z_{m10} &= E \qquad Z_{m4} + Z_{m9} = F \qquad Z_{m2} + Z_{m8} = G \\ Z_{m2} + Z_{m7} &= H \qquad Z_{m1} + Z_{m6} = J \qquad Z_{m2} + Z_{m5} = K \end{aligned}$$

we obtain from the equations above, the following relationships :—

$$\frac{I_1}{I_0} = \frac{CDH - CKF - DEJ + EGK + FGJ - HC^2}{CF^2 + BG^2 + DE^2 - BCD - 2EFG} = A_1 \underline{\alpha_1}$$

$$\frac{I_2}{I_0} = \frac{BDJ - BGK - DEH + EFK + FGH - JF^2}{CF^2 + BG^2 + DE^2 - BCD - 2EFG} = A_2 \underline{\alpha_2}$$

$$\frac{I_3}{I_0} = \frac{BCK - BGJ - CFH + EFJ + EGH - EK^2}{CF^2 + BG^2 + DE^2 - BCD - 2EFG} = A_3 \underline{\alpha_3}$$

These are shown plotted against reflector reactances in Fig. 13.

Defining the angle θ as shown in Fig. 12(b) ; and β_1, β_2, β_3, by the equations

$$\begin{aligned} \beta_1 &= \alpha_1 - (2\pi d/\lambda) \cos\theta \\ \beta_2 &= \alpha_2 - (2\pi d/\lambda) \cos\theta \\ \beta_3 &= \alpha_3 - (2\pi d/\lambda) \cos\theta \end{aligned}$$

we find that the field strength at P is proportional to $\cos\theta(1 + A_1 \underline{\beta_1} + A_2 \underline{\beta_2} + A_3 \underline{\beta_3})$.

The amplitude is given by

$$\cos\theta\sqrt{(1 + \Sigma A \cos\beta)^2 + (\Sigma A \sin\beta)^2}$$

Theoretical polar diagrams evaluated by means of this expression are shown in Fig. 14. In general a close similarity exists between these polar diagrams and those of the two-stack single-reflector systems but with this difference : there is a marked increase in bandwidth in the case of the triplet system.

The ratio of the maximum field strength of the largest backward lobe to that of the forward lobe

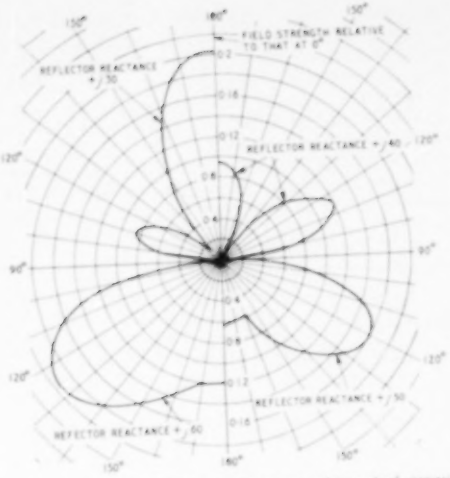
FIELD STRENGTH RELATIVE TO THAT AT 0°
REFLECTOR REACTANCE +j 30
REFLECTOR REACTANCE +j 40
REFLECTOR REACTANCE +j 50
REFLECTOR REACTANCE +j 60

Fig. 10. *Theoretical polar diagrams of two-stack array with single reflectors.*

is less than 0.17 from $j80$ to $j140$ [see Figs. 13(d) and 14]. For the two-stack single-reflector system this ratio is only less than 0.17 between the limits of reflector reactance $j35$ and $j50$, while between the limits $j30$ and $j60$ it is less than 0.2.

The 'clover leaf' patterns are almost identical in the two cases with a back-to-front ratio of about 0.12, the main difference being that a reflector reactance equal to $+ j90$ ohms is required for the triplet array whereas $j40$ ohms is required for the array using single reflectors.

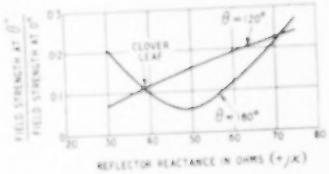

Fig. 11. *Back-to-front ratio two-stack single-reflector array.*

5. Reactance and Length of a Parasitic Element

The question at once arises as to the connection between reflector length and its reactance, for in the construction of an array it is the physical

dimensions that need to be known. Now reflector length can be adjusted experimentally until the polar diagram of minimum radiation at 180° is realized. A further shortening can then be effected so as to obtain the clover-leaf pattern. This change of length, therefore, is equivalent to the difference in reactance given by theory between these two well-defined points. Hence it is possible to work out the non-reactive length of a reflector and also the change of reactance per 1% change in length (referred to $\lambda/2$ for convenience) of reflector.

The results of performing this calculation for the two-stack arrays under consideration is given in Table I.

TABLE I

Type of Two-Stack Array	Non-reactive length	Change of reactance per 1% change in length
Triplet reflector . .	0.48λ	14.9 ohms
Single reflector . .	0.455λ	7.2 ohms

In the arrays used experimentally all elements were made of 200 lb/mile copper wire having a diameter of 0.11 inch. This gives $Z_0 = 800$ approximately for the 6-metre band.

If X is the reactance of an element measured at its centre then $\frac{dX}{d(1/\lambda)} = \pi Z_0$,

whence a reactance change of 12.5 ohms per 1% change in reflector length is quickly evaluated for the case under consideration.

Clearly this reactance change should be a constant for arrays of the same Z_0. The inconsistency of the results may be in part due to the limitations of the sinosoidal theory which assumes that the phase of the current along each wire is constant, a state of affairs which certainly is not true. Good agree-

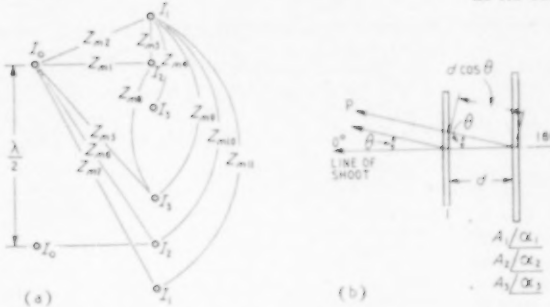

Fig. 12 (above). *Elevation (a) and plan (b) of two-stack triplet-reflector array.*

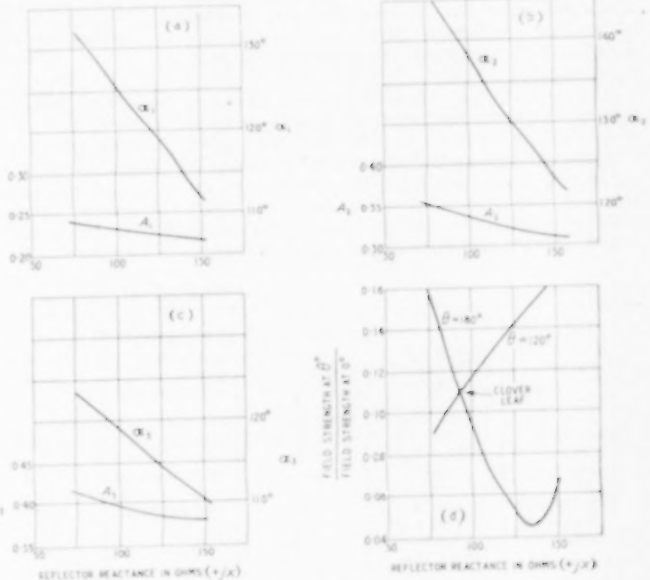

Fig. 13 (right). *Current ratios A_1, A_2 and A_3 and phase angles α_1, α_2, α_3 of two-stack triplet-reflector array are given in (a), (b) and (c), with the back-to-front ratio in (d).*

It is interesting to compare these results with those given by ordinary line theory. Z_0, the characteristic impedance of a parasitic element, can be expressed by the formula :—

$$Z_0 = 120\left[\log_e\left(\frac{l}{a}\right) - 1\right]$$

where l = length of reflector

and a = radius of reflector wire or tube.[2]

ment however, between theory and practice in this respect has been obtained in the case of aerials made of tubing; i.e., having a much lower value of Z_0.[3]

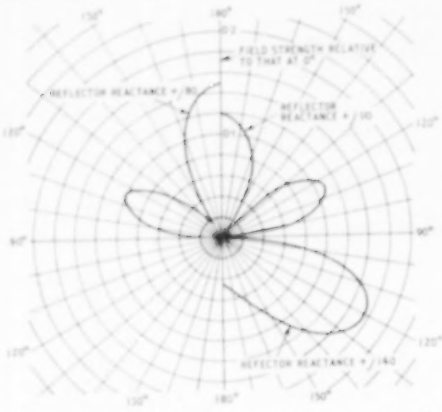

Fig. 14. *Theoretical polar diagram of two-stack triplet-reflector array.*

6. Experimental Work

A two-stack triplet-reflector array constructed from 200 lb/mile copper wire was rigged at a mean height of about 30ft, on a light wooden tower erected on a flat site. Effective shortening of the reflectors due to insulators was avoided by bending small hooks at the ends of the wire elements and tensioning them in position with fine string.

A five-watt c.w. transmitter was used to energize the array and reception was carried out at points round a circle 80 yards in radius, using a horizontal half-wave dipole mounted on a pole about 0.6λ above ground. The array was constructed for a nominal wavelength of 5.9 m, reflector length was 3.01 m($\equiv$0.51λ), and horizontal spacing between dipoles and reflectors was 0.17λ. Vertical spacing between dipoles was, of course, $\lambda/2$.

It will be noticed here that the approximation was made of comparing these experimental polar

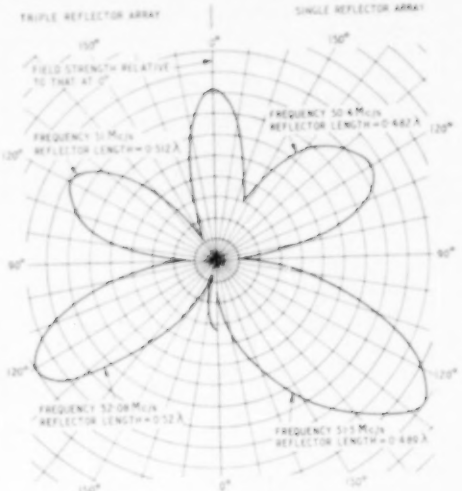
TRIPLE REFLECTOR ARRAY
SINGLE REFLECTOR ARRAY
FIELD STRENGTH RELATIVE TO THAT AT 0°
FREQUENCY 50.4 Mc/s
REFLECTOR LENGTH = 0.482 λ
FREQUENCY 51 Mc/s
REFLECTOR LENGTH = 0.512 λ
FREQUENCY 52.08 Mc/s
REFLECTOR LENGTH = 0.52 λ
FREQUENCY 50.5 Mc/s
REFLECTOR LENGTH = 0.489 λ

Fig. 15. *Experimental polar diagrams, the arrays being at a mean height of 3λ/2 above ground. For the single reflector the reflector length was 0.487λ and other dimensions were as in Figs. 1 and 3 when λ = 5.9 m. For the triplet, when λ = 5.9 m, the reflector length was 0.51λ and the other dimensions were as in Figs. 2 and 12.*

diagrams with those calculated for the aerial system in free space. This is done to avoid unwieldy complications to the theory, and should not involve big errors in practice because

(a) the height of the experimental aerial system above earth was sufficient to render mutual coupling with its images relatively small,

TABLE II

Type of Array	Frequency of Polar Diagram (Mc/s)	Back (180°) to Front (0°) Ratio of Signal Strength		Reflector Length	Reflector Reactance Theoretical (Ω)	Change in React. per 1% change in length (based on λ/2) (Ω)
		Theoretical	Experimental			
Two-stack Triplet Reflector	51 (clover leaf)	0.122	0.17 approx.	0.512λ	+ j90	14.9
	52.08	0.044	Not measured	0.52λ	+ j120	
Two-stack Single Reflector	50.55 (clover leaf)	0.123	0.17 approx.	0.482λ	+ j40	7.2
	51.5 (min. 180°)	0.062	0.04	0.489λ	+ j50	

(b) the distance between transmitting and receiving aerials was sufficient for expecting reasonable correspondence between the horizontal polar diagrams measured and those calculated for free-space conditions.

Polar diagrams in a back quadrant were measured at :

(1) 52.08 Mc/s, which was expected to represent minimum 180° radiation, but from the presence of the small lobe at 180° (see Fig. 15) appears to be 0.5 Mc/s low in frequency;

(2) 51 Mc/s, which is the clover-leaf pattern.

A measurement of back-to-front ratio was also made at this frequency.

Results are given in Table II together with those obtained for the single-reflector two-stack system by way of comparison ; the polar diagrams are plotted in Fig. 15. From these results the bandwidth comparison given in Table III for two-stack arrays at the nominal wavelength of 5.9 metres was calculated.

TABLE III

Type of Array	Frequency Shift (Mc/s)	Altering polar Diagram pattern		Equivalent change in Reflector length to give same alterations (cm)
		From	To	
Two-stack Triplet Reflector	1.5	Clover leaf	Min. 180° radiation	9
Two-stack Single Reflector	1	Clover leaf	Min. 180° radiation	5.9

7. Conclusion

Theory would indicate that, for a given standard of performance, the bandwidth of the triplet system should be well over twice that of the ordinary single reflector system, but the experimental evidence available did not bear this out. Whereas, the change in reflector reactance per 1% change in length was, in the case of the triplet-reflector array, not far removed from the theoretical value of 12.5, it was, in the case of the single-reflector system, little more than half this theoretical value. Further experimental work might throw light on this discrepancy but as it stands the triplet shows an increase in bandwidth of the order of 50% over the single-reflector system, or, putting in figures at $\lambda = 6$ m, a workable band of 3 Mc/s as opposed to 2 Mc/s for a two-stack array of the single-reflector type.

8. Acknowledgments

The work which forms the subject matter of this paper was carried out by the author at the Telecommunications Research Establishment as part of the wartime research programme of the Ministry of Aircraft Production. Acknowledgments are due to the Chief Scientist, Ministry of Supply, for permission to publish the paper. Crown copyright reserved. Reproduced with the permission of the Controller of H.M. Stationery Office.

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FIXED H-ADCOCK DIRECTION FINDER FOR V.H.F.

By B. G. Pressey, Ph.D., M.Sc.(Eng.), A.M.I.E.E. and G. E. Ashwell, B.Sc.

(Communication from the National Physical Laboratory)

SUMMARY.—The paper describes an investigation into the practicability of the fixed type of H-Adcock direction finder for use at very high frequencies (30-100 Mc/s) under conditions in which the aerial system is remote from the operator. The experimental equipment consisted essentially of two crossed H-Adcock aerials mounted on a wooden tower 10 m high. The aerial system was connected by r.f. transmission lines to a goniometer and receiver situated in a hut near the base of the tower. By making adjustments to the length of the transmission lines and their point of connection to the aerial feeders a high instrumental accuracy was obtained on signals of mixed as well as vertical polarization. The sensitivity was such that bearings with a silent swing of $\pm 5^\circ$ could be taken on field strengths varying between 0.5 and 14 μV/m over the frequency range.

1. Introduction

FOR many applications of v.h.f. direction finders it is important that the aerial system should be situated at some distance from the operator. One particular group of applications is that in which the aerials must be erected at an appreciable height above ground level in order that they shall be clear of obstacles on the site which would seriously impair their performance. A direction finder on board ship is a specific example. It was the use of the H-Adcock system in such cases which was the object of the investigations described in this paper.

At the time when this work was started (June 1942) the Adcock systems which had been developed for this frequency range were of the rotating type.[1] For several reasons, such as the serious effect of wind resistance and moment of inertia of the aerials on the performance of the remote driving arrangements, this type of direction finder was considered unsuitable for the present application. An investigation was therefore made into the practicability of the fixed aerial type erected in an elevated position at a distance of 20 m or more from the operating point.

The experimental system described consisted essentially of two crossed H-Adcock aerials mounted on the top of a 10-m wooden tower. The aerials were connected by r.f. transmission lines to a goniometer and receiver situated in a hut near the base of the tower. The frequency range was limited by the receiver to 27-85 Mc/s.

2. Description

2.1 *Aerial System.*

The aerial system was supported on a wooden framework which was mounted in the centre of a platform at the top of the tower. Each of the two aerial pairs consisted of two vertical dipoles 1.6-m long, spaced 1-m apart and interconnected by a screened horizontal feeder. Dipoles having a cross-sectional area large compared with their length were used so as to reduce the variation of their impedance with frequency. The diameter was 9 cm giving a ratio of diameter to length of 0.055. The resonant frequency was 73 Mc/s. The horizontal feeders consisted of two concentric r.f. cables of 70-Ω characteristic impedance and 2.6 m in length. When the system was first erected the cables were made as short as possible (1.2 m) but it was found that the natural resonance of the aerial pair occurred within the operating frequency range. Satisfactory operation was impossible in the region of this resonance and it was necessary to lengthen the feeders so that the resonance occurred outside the range at a frequency of approximately 27 Mc/s.

The transmission lines from the aerial system to the receiver consisted, like the horizontal feeders, of a pair of 70-Ω concentric cables and were 23 m long. The lines were connected directly to the mid-points of the feeders in the manner shown in Fig. 1. In order to facilitate the adjustment of the actual point of connection when balancing the aerial system, the central portion of the feeders consisted of two thick brass wires bent into semi-circles and mounted on either side of a block of insulating material as shown in Fig. 2. The diameter and spacing of the wires were so chosen that the characteristic impedance of the feeders was maintained. Connection of the lines to these wires was effected through two rotatable contact arms. The lower ends of the lines terminated on the field-coil terminals of the goniometer and a 200-Ω resistor, centre-tapped to earth, was connected in parallel for the purpose of damping line resonances.

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The cables used for the feeders and lines had lead sheaths which were carefully bonded to each other and earthed at the foot of the tower.

2.2 Goniometer.

The goniometer was of the inductive type with electrostatically-screened field coils and a relatively high coupling factor. A full description of the instrument and of its performance is given elsewhere.[2]

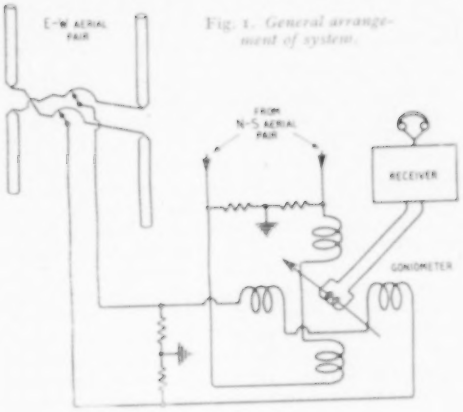

Fig. 1. *General arrangement of system.*

2.3 Receiver.

The receiver was a commercial type, the Plessey RL85, covering a frequency band of 27–85 Mc/s in three ranges. Certain improvements to the screening were found to be necessary in order to reduce stray pick-up. The goniometer search coil was connected by a length of twin screened cable ($Z_0 = 100\ \Omega$) to the range switch in the grid circuit of the first r.f. stage and was made as short as possible so as to avoid any resonance of the search-coil circuit within the operational frequency band.

3. Aerial Balance Adjustments

Although the aerial system was constructed to be mechanically symmetrical and the transmission lines were cut to equal lengths, the electric symmetry and equality were not correspondingly good. It was necessary, therefore, to make adjustments to the aerials and the lines in order that the system should give the optimum performance. The adjustments may be conveniently divided into two sections, those for the reduction of 'antenna effect' and those for the reduction of polarization error.

3.1 Reduction of Antenna Effect.

Measurement of the antenna effect of each aerial pair was made with the aid of a transmitter with a short vertical aerial. The transmitter was set up on a portable tower at a distance of about 5 m from the direction finder and at the same height. This arrangement was adopted so as to minimize site errors which were appreciable when the transmitter was placed on the ground at a distance of 40 m. The transmitter aerial was set carefully on a line normal to the plane of the aerial pair under examination. The error in bearing was measured over the frequency range at intervals of 1 Mc/s and plotted against frequency. A typical curve for a well-balanced system is shown in Fig. 3. The error curve has two principal components, one varying slowly with frequency, as indicated by the broken line, and the other having the form of rapid fluctuations with a period of between 2 and 4 Mc/s. The former component is due mainly to asymmetry of the aerial pair and the latter to unbalance in the transmission line.

The aerial adjustments for the reduction of antenna effect were made in two stages; first, to the lengths of the line cables, and secondly to the position of the point of connection of the lines to the feeders.

When the direction finder was originally set up screened twin cables were used for the line and the error curve fluctuations were found to be of the order of 5° in 1 Mc/s. By the addition of balancing impedances between the conductors and the sheath of the cable at its lower end it was possible to reduce this error to reasonable proportions at a particular frequency but the

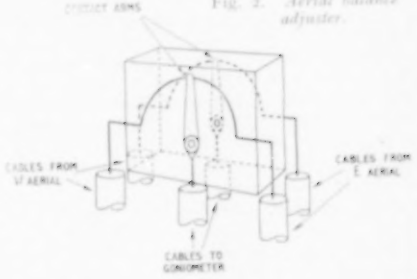

Fig. 2. *Aerial balance adjuster.*

balance did not hold at other frequencies. The substitution of two concentric cables in place of the twin cable enabled the relative lengths of the two cables to be adjusted by cutting until the amplitudes of the error fluctuations were reduced over the whole frequency range to the order of those shown in Fig. 3.

Aerial balance was achieved by moving the two line tapping points together along the feeders until the electrical centre point was found. The adjustment was most critical at

frequencies near the resonance of the aerial pair (27 Mc/s), a movement of 1 mm causing a change of about 1° in the bearing error. Thus adjustment for minimum error on a frequency in this region produced the most satisfactory balance throughout the range.

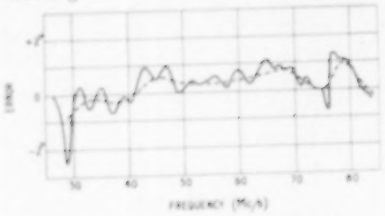

Fig. 3. *Variation of aerial balance with frequency for N–S aerial pair. Full line = measured errors; broken line = mean errors curve.*

3.2 *Polarization Error*

In order to minimize the horizontal pick-up of the aerials it is necessary that the aerials be electrically symmetrical about the horizontal plane containing the feeders. This symmetry was obtained by further adjustment of the connection points of the lines to the feeders in the same manner as has been used on h.f. direction finders.[3]

As in the case of the antenna-effect adjustments the aerial pairs were balanced in turn using a local transmitter sited on the line normal to the plane of the aerial pair. The transmitter had a dipole aerial capable of rotation in the vertical plane parallel to that of the d.f. aerials. Thus it was possible to introduce a horizontally-polarized component into the incident wave and measure the resultant change in bearing and swing. The dipole aerial was situated at a height of 4 m above the ground and at a distance of 46 m from the direction finder. Measurements over the frequency range showed peak values at 73 Mc/s (dipole resonance) and at 27 Mc/s (aerial-pair resonance), the latter being considerably greater. Accordingly, the two connection points of the lines were moved in opposite directions along the feeders, their mean position (controlling the antenna effect) remaining the same, until the error and swing at this frequency reached a minimum value. The adjustment was as critical as that for antenna effect, the error changing at a rate of approximately 2°/mm.

When the adjustments had been completed for both antenna effect and polarization error the contact arms were soldered in position to prevent further movement and to ensure a good contact.

3.3 *Balance Between Aerial Pairs*

When each aerial pair had been separately balanced it was necessary to make a further adjustment so that the terminal voltages of the two transmission lines should bear the correct relation to each other. For this purpose the test transmitter was set up as for the antenna-effect adjustments but at 45° azimuth. Short lengths were then cut off both cables of one of the lines until the rapid fluctuations in the bearing error/frequency curve reached minimum amplitude.

4. Performance Measurements

When all adjustments to the aerials and lines were finished, complete measurements of the instrumental accuracy and sensitivity were made.

4.1 *Instrumental Accuracy*

A bearing calibration of the direction finder for vertically-polarized waves was made at frequencies of 31, 45, 60 and 80 Mc/s. The test transmitter was set up at a distance of 5 m in the same manner as for the balance adjustments and the d.f. aerial system was rotated in steps of approximately 11°, readings of bearing being taken at each position. For the reason explained below, it was decided to eliminate the effects of the rapid fluctuations of bearing with frequency, by making measurements on a group of three frequencies separated by 1 Mc/s and centred on each of the frequencies stated above. The mean value of the three readings was used in plotting the error curves, two of which, those for 31 and 60 Mc/s, are shown in Fig. 4.

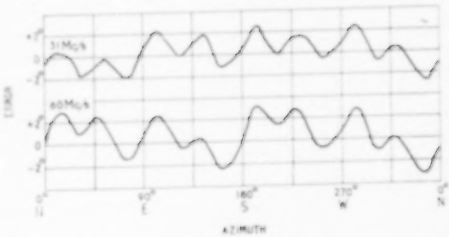

Fig. 4. *Measured calibration curves.*

As stated in Section 3.1, the reason for locating the transmitter very close to the direction finder was the elimination of site error. It is possible, however, to introduce calibration errors due to the proximity of the transmitter, the magnitude of which have been examined by Ross.[4] It can be shown that in the present case the principal error is octantal having a maximum value of about ± 0.2° over the whole frequency range.

The total errors indicated by the curves of Fig 4 are the sum of the errors due to aerial spacing, aerial unbalance, line asymmetry and goniometer imperfections. The maximum value of the octantal-spacing error varied from 0.3° at 30 Mc/s to 1.8° at 85 Mc/s and the goniometer

errors at 31 and 60 Mc/s are shown by the full-line curves of Fig. 5.

Subsequent to the making of these calibrations, considerable improvements in goniometer performance were effected as indicated by the broken line curves of Fig. 5. Unfortunately it was not possible to repeat the calibrations, in order to determine whether there was an overall improvement, but an examination of Figs. 4 and 5 will show that, had this been possible, smoother calibration curves would have been obtained.

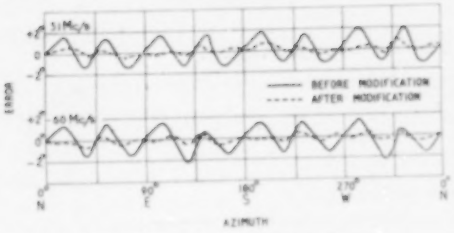

Fig. 5. *Goniometer error curves.*

The method which has been used in obtaining the above calibrations effectively eliminates the rapid fluctuations of error with frequency which are evident in Fig. 3. Examination of this curve shows that in general it would be impracticable to correct for this error on account of the large number of correction curves which would be required. Under special conditions of operation, such as operation over a limited sector of frequency range, however, the use of complete correction curves might be practicable. But, for normal operation of the direction finder, the presence of these rapidly fluctuating errors must be accepted and correction made only for the slowly varying errors. From measurements made during the course of development it is possible to estimate the limit of accuracy due to this uncorrected error for various types of transmission line. Thus, Table 1 shows the highest accuracy which can be expected under various line conditions.

TABLE 1

Transmission-Line Condition	Limiting Accuracy
Twin cable, mechanically equal lengths	$\pm 6°$
Pair of concentric cables, mechanically equal lengths	$\pm 4°$
Pair of concentric cables, cut to electrically equal lengths before installation	$\pm 2°$
Pair of concentric cables, balanced in situ	$\pm 1°$

The error of the direction finder for a wave containing a horizontally-polarized component was measured in the manner described in Section 3.2 with the transmitter placed in line with each aerial pair in turn. From the measured bearing error and swing the total polarization error (i.e., the maximum error obtained when the horizontal and vertical components have the appropriate phase relations) was calculated in the manner already described in a previous paper by one of the authors.[5]

The set up of the transmitter was such that the ratio of the horizontal and vertical components of the field at the d.f. aerials, which had been measured prior to the installation of the aerial system, varied between 5 and 6 over the frequency range. In order to express the results in a rational manner the measured errors were converted to the values which would be obtained with a wave arriving in a horizontal direction and having a ratio of horizontal to vertical field components of unity. The total error under these conditions and with the transmitter in line with one aerial pair is shown in Fig. 6. The small peak value at 72 Mc/s and the high values below 30 Mc/s are associated with the dipole resonance and the aerial-pair resonance respectively. The average error of 0.4° corresponds to a ratio of horizontal to vertical pick-up factors of 0.007.

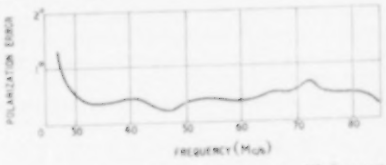

Fig. 6. *Variation of polarization error with frequency for E–W aerial pair. Angle of polarization = 45°; angle of incidence = 90° to vertical.*

4.2 *Sensitivity*

The overall sensitivity of the system is expressed in terms of the field strength of a c.w. signal on which a bearing could be taken with a silent swing of $\pm 5°$, an audible output being obtained by using the internal beat oscillator of the receiver. A transmitter with a vertical aerial provided a signal of known field strength at the d.f. aerials. Only one aerial pair was used for this measurement, the transmitter being set up in line with it, and by connecting a signal generator to the unused goniometer field coil it was possible to inject into the receiver a voltage equal to that produced by the radiated field. By reducing the generator output until just sufficient for a bearing with a $\pm 5°$ swing to be taken, the level of the minimum field strength

relative to that actually incident on the aerials was determined. The measurements were made at intervals over the frequency range and the results are plotted in Fig. 7. The large increase in sensitivity at 70 Mc/s is due chiefly to the improved impedance match between dipoles and feeders at the resonance of the former.

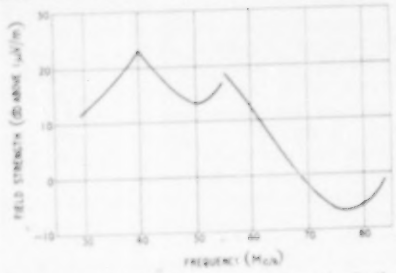

Fig. 7. *Field strength of c.w. signal for ± 5° silent swing.*

5. Operational Performance

The equipment has been operated on routine observations of signals from c.w. and pulse transmitters at distance of 6 to 100 km with satisfactory results.

As the main purpose of the investigation was the development of the equipment and the determination of its performance as an instrument, no great care was taken in choosing a site that would be suitable for observations on distant transmitters. Thus it was found that when taking bearings in connection with the above observations appreciable site errors were present. For instance, the bearing of a transmitter 6 km distant showed a variation with frequency of 2°–3° and varied at a rate of up to 2½° for a frequency change of 0.1 Mc/s. With transmissions on 47 Mc/s from a distance of 35 km the bearings showed a variation of 9° for a frequency change of 1.0 Mc/s. Even on a good site, the errors due to the site will rarely be less than about 1° to 2°, so that the instrumental accuracy which has been achieved is quite adequate in practice.

6. Conclusions

The development of the direction finder and the investigation of its performance has shown that the fixed type of H-Adcock system may be considered a practical v.h.f. system capable of operation over a wide frequency range with a high order of accuracy. It can be made simple in operation having only receiver adjustments apart from the goniometer operation. The initial balancing of the aerial system and transmission lines is rather critical if the greatest accuracy is required but if a maximum probable error of the order of 2° is satisfactory the balancing processes can be considerably simplified.

7. Acknowledgment

The work described above was carried out as part of the programme of the Radio Research Board. This paper is published by permission of the Director of the National Physical Laboratory, and the Director of Radio Research of the Department of Scientific and Industrial Research.

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INSTITUTION OF ELECTRICAL ENGINEERS

The following meetings will be held at the I.E.E., Victoria Embankment, Savoy Place, London, W.C.2, commencing at 5.30 :—

Feb. 13th. Discussion meeting on "Interference Suppression," opened by E. M. Lee.

Feb. 15th. "Ground-Wave Propagation over an Imhogeneous Smooth Earth—Experimental Evidence and Practical Implications," G. Millington and G. A. Isted.

Feb. 27th. Discussion Meeting on "Mobile Radio Power Packs," opened by Air Commodore R. L. Phillips, O.B.E.

Mar. 6th. Discussion Meeting on "The Place of High-Frequency Heating in Industry," opened by C. E. Eadon-Clarke.

Mar. 7th. "The Fundamental Limitations of Second-Harmonic Type of Magnetic Modulator as applied to the Amplification of Small D.C. Signals," by Professor F. C. Williams, O.B.E. and S. W. Noble, and "A New Theory of the Magnetic Amplifier," by A. G. Milnes.

BRITISH INSTITUTION OF RADIO ENGINEERS

During February the following meetings will be held :—

London Section, 3rd Mar., at 6.30 p.m., at the London School of Hygiene & Tropical Medicine, Keppel St., W.C.1. : "Travelling Wave Tubes," by R. L. Kompfner.

West Midlands Section, 1st Mar., at 7 p.m., at the Wolverhampton & Staffordshire Technical College, Wulfruna St., Wolverhampton : "Electronics and the Brain," by H. W. Shipton.

North Eastern Section, 15th Feb., at 6 p.m., at Neville Hall, Westgate Rd., Newcastle-on-Tyne : "Single Sideband Systems Applied to Long Range Wireless," by Major S. R. Rickman.

ELECTRON TRANSIT TIME

Effect on Negative-Grid Oscillators

By S. K. Chatterjee, M.Sc., M.I.R.E., and B. V. Sreekantan, M.Sc., A.M.I.R.E.

(Department of Electrical Communication Engineering, Indian Institute of Science, Bangalore, India)

SUMMARY.—Electron transit-time effects on the output and efficiency of some triodes having different geometrical structures have been studied. The results indicate that for valves 833-A, 834, 304-B, the limit of oscillation is reached when the period of oscillation is nearly four times the total time of transit of the electrons from cathode to anode, whereas for valves 316-A, 955, the limiting frequency is reached when the period of oscillation is nearly three times the total time of flight. The transit-time expression given by Gavin[1] has been modified by taking into consideration the alternating voltages on the anode and grid and the negative bias. The wavelengths at which the efficiencies of valves drop by 10% and also down to 10%, have been found and the results compared with values obtained theoretically from the expression given by Gavin.[1]

Introduction

THE highest frequency at which a valve can generate spontaneous oscillation is restricted by the time of flight of electrons from cathode to anode, the interelectrode capacitances and the lead inductances within the valve. When the time required by an electron to travel from cathode to anode becomes comparable with the period of a cycle of oscillation to be generated, the valve performance is modified in a fundamental way and the entire behaviour needs clarification in the light of basic considerations. Electron transit-time increases the effective grid-cathode conductance proportional to the square of the frequency and hence decreases the effective Q of the resonant circuit. At these frequencies the phase relation between anode current and grid voltage is no longer that existing at low frequencies. So the phase angle of the mutual conductance is also affected. A considerable amount of work has been done by Scheibe,[2] Llewellyn,[3] North,[4] Benham,[5] and McPetrie,[6] to evaluate the effect of electron transit time on the assumption that a steady-state space-charge distribution has been established and the high-frequency signal is merely a perturbation, small in comparison with the steady-state component. As remarked by Llewellyn[7] such investigations are inadequate for the case of large signals encountered in power amplifiers and oscillators operated in the class C condition. Expressions which can be used to analyse the type of signals involved in class C operation, have been derived by Chao-Chan-Wang.[8]

The time of electron flight for valves having plane and cylindrical electrodes has been calculated by Fortescue[9] and Kompfner.[10] Assuming that for $r_a/r_g < 3$ and $r_g/r_k < 2$, where r_a, r_g, and r_k represent radii of anode, grid and cathode respectively, plane electrodes may be considered as giving values of transit time which agree within 12% with the values for cylindrical electrodes. Gavin[1] has shown that the time of flight from cathode to anode is given by the following expression

$$\tau = \frac{\sqrt{\mu}}{5.93 \times 10^7 \sqrt{E_{HT}}}\left(3d_{kg} + \frac{2d_{ga}}{1+\sqrt{\mu}}\right) \quad (1)$$

involving cathode to grid distance d_{kg}, grid to anode distance d_{ga}, amplification factor μ and anode voltage E_{HT}. In the above expression, the effective grid voltage has been taken to be E_{HT}/μ. But, in the case of an oscillator working under class C conditions, the effective voltage at the grid when the electron that left the cathode just when the grid started on its positive cycle reaches the grid plane is shown in the Appendix to be

$$(-E_{GB}+E_{HT}/\mu)+(E_g-E_a/\mu)\cos(\theta_g/2-\theta_{kg})$$

where

E_{HT}, E_{GB} = steady anode and grid supply voltages

E_a, E_g = amplitudes of alternating components of anode and grid voltages

e_a, e_g = instantaneous anode and grid voltages.

Hence the cathode to grid time of flight τ_{kg} can be shown to be

$$\tau_{kg} = \frac{3d_{kg}\sqrt{(m/2e)}}{\sqrt{[(-E_{GB}+E_{HT}/\mu)+(E_g-E_a/\mu)\cos(\theta_{kg}-\theta_g/2)]}} \quad \ldots\ \ldots\ \ldots\ \ldots \quad (2)$$

where m/e is the ratio of mass to the charge of an electron.

Assuming negligible space-charge effect in the grid-anode space, the transit time τ_{ga} for an electron which leaves the grid when the grid is at its positive peak is given by the following expression. It may be mentioned that the electron which left the cathode at the instant the grid

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started on its positive cycle reaches the grid at the instant that the grid voltage attains its positive peak. If $E_g > E_{gt}/\mu$ (Appendix)

$$\tau_{gg} = \frac{2}{B \times 5.93 \times 10^7}\{\sqrt{(A + B \cdot d_{gg})} - \sqrt{A}\}$$

where

$$A = (E_{gt}/\mu - E_{ga}) + (E_g - E_a/\mu)$$

$$B = \left[\{E_{gt} - (E_{gt}/\mu - E_{ga})\} - \{E_g + (E_g - E_a/\mu)\}\right] d_{gg}$$

The total cathode to anode time of flight τ is, therefore, given by

$$\tau = \tau_{kg} + \tau_{gg} = \frac{\sqrt{\mu}}{5.93 \times 10^7 \sqrt{E_{gt}}} \cdot \left(3d_{kg} + \frac{2d_{gg}}{1 + \sqrt{\mu}}\right)$$

when we neglect the alternating voltages on the electrodes and the bias on the grid. This is the same expression as that given by Gavin.[1]

The expression (2) is not readily useful for evaluating the transit time as it involves θ_{kg} and θ_{gg} but certain important conclusions can be drawn. When the amplitude of the alternating voltage on the grid is $1/\mu$ times the amplitude of the alternating anode voltage, then for an electron which leaves the cathode at the instant at which the grid starts becoming positive, the transit time τ_{kg} is independent of the amplitudes of the alternating voltages on the grid and anode. Another important feature is that when there is no lag (i.e., $\theta_{kg} = \theta_g/2$) the transit time τ_{kg} is reduced to a minimum if $E_g > E_{gt}/\mu$ as given by the following expression

$$\tau_{kg} = \frac{3d_{kg} \cdot \sqrt{(m/2e)}}{\sqrt{\left(-E_{ga} + \frac{E_{gt}}{\mu}\right) + \left(E_g - \frac{E_a}{\mu}\right)}} = \frac{3d_{kg} \cdot \sqrt{(m/2e)}}{\sqrt{\frac{E_{g(min)}}{\mu} + E_{g(max)}}} = \frac{3d_{kg}}{5.93 \times 10^7 \sqrt{A}}$$

Investigations by Gavin[1] show that the limiting frequency of oscillation is reached when the period of oscillation generated becomes equal to twice that of electron transit, whereas the result of Wagener[11] shows that the limiting frequency is reached when the cathode-to-grid transit time is one-sixth the period of oscillation. This difference is perhaps due to the difference in structures of the valves investigated. Modern technique has made it possible to evolve various valve structures in an attempt to extend the frequency range and useful output. So it has been thought worthwhile to study experimentally the effect of transit time on the output and efficiency and also to find the limiting frequency of oscillation of some of the available valves having different electrode structures.

Results and Discussion

Variation of power output with frequency for different valves (Fig. 1) shows the inevitable reduction in output with increasing frequency. Using dimensional analysis, Lehmann and Vallarino[12] have shown that power output W is given by the following expression.

$$W = H_4 (J^5/f^{10})$$

where H_4 is a constant, J denotes current density and f is the frequency in megacycles per second. The above expression signifies that for the same type of cathode, the power available from a valve drops very rapidly with increasing frequency. Though the above relation does not quantitatively represent the practical cases it clearly indicates that, to maintain considerable output, the current density $J = \rho v$ per unit area must be maintained at a high value, by making the charge density ρ and/or the speed of electrons v high. Two more facts are significant; first, the curves having the same general shape, it may be said that the same factors are responsible for limiting the output; secondly, that curves for some of the valves approach a limiting curve, shown by the dotted line. This common asymptote indicates that some common physical limitation is present. Moreover, greater power is available from larger valves at lower frequencies, but decay of output starts at lower frequencies for these valves.

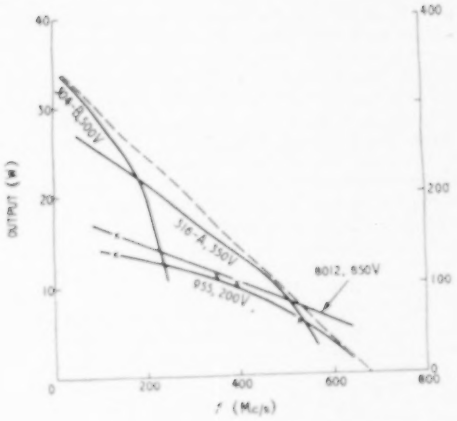

Fig. 1. *Variation of output with frequency for different types of valve. The dotted line shows that the curves approach a limiting condition.*

The inherent losses in the functioning of vacuum valves results from the degradation of kinetic energy of the electrons into heat at impact on the electrode surface. Consequently, efficiency η

depends on the motion of electrons. Efficiency being a dimensionless parameter, it has been shown[12] that $fd/\sqrt{V}$ (where d is the cathode-anode distance in cm and V is the anode potential in volts) can be represented by a dimensionless parameter, so the expression for η can be written as follows

$$\eta = u(\phi, \theta_1, \theta_2 \ldots \theta_n)$$

where ϕ is a dimensionless parameter and is taken to be equal to $fd/\sqrt{V}$ and the θ terms are dimensionless parameters, containing ϕ and describing conditions of adjustment of the oscillating circuit. If the θ terms are so chosen that η becomes a maximum η_{max} then $\eta_{max} = u(\phi)$, where one value of ϕ corresponds to a unique value of η_{max}. The variation of η with $fd/\sqrt{V}$ (Fig. 2) indicates the maximum efficiency realisable under specific conditions.

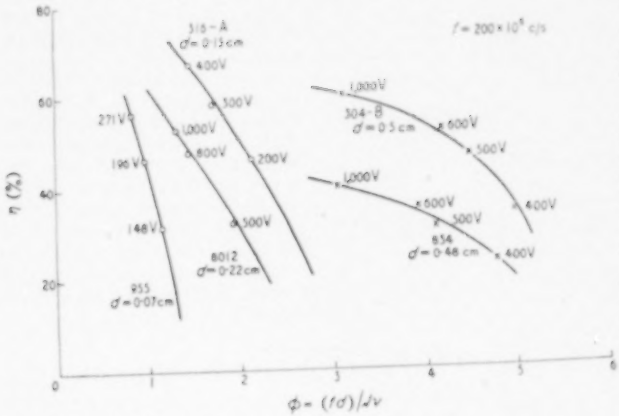

Fig. 2. (Above) *Maximum efficiency realizable under specific conditions*

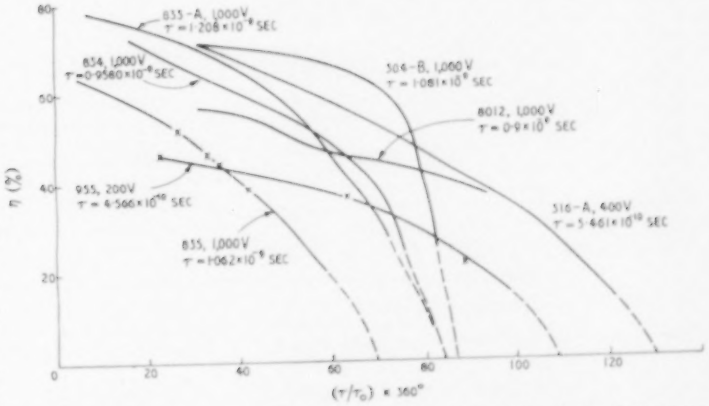

Fig. 3 (right). *Efficiency versus transit angle. The limit of oscillation is found from the point where the dotted curves meet the horizontal axis.*

To study the effect of transit time Equ. (1) has been employed. Substituting the proper values for d_{xa} and d_{kx}, Equ. (1) can be written as follows

$$\tau_{316\text{-}A} = \frac{2.54 \times 0.255}{5.93 \times 10^7 \sqrt{E_{HT}}}$$

$$\tau_{8012} = \frac{4.243 \times 0.398}{5.93 \times 10^7 \sqrt{E_{HT}}}$$

$$\tau_{955} = \frac{5 \times 0.0766}{5.93 \times 10^7 \sqrt{E_{HT}}}$$

where the subscript to τ indicates the valve type number.

If τ_0 represents the period of oscillation, then the transit angle, $\tau/\tau_0 \times 360°$, at which a particular valve reaches its limit of oscillation is shown by extrapolating the curves of efficiency η v. transit angle (Fig. 3). The transit angle corresponding to the frequency limit of each valve is given in Table I.

It will be observed that for valves 833-A, 304-B and 834 the limiting frequency of oscillation is reached when the period of oscillation is nearly four times the time of transit of electrons from cathode to anode. In the case of 316-A and 955 the transit time corresponding to the limiting frequency is nearly one-third the period of oscillation.

The effect of transit time on efficiency can further be studied by calculating the differences in efficiencies η_1 and η_2 at a long wavelength and at a particular ultra-short wavelength. The difference in efficiencies for class C oscillators is given[1] by the following expression

$$\eta_1 - \eta_2 = \frac{\sqrt{2}}{\pi}\left(\frac{E}{E_{HT}}\right)\omega_0^2\tau^2 \quad \ldots \quad \ldots \quad (3)$$

where E/E_{HT} denotes the ratio of peak alternating voltage to the mean voltage of the anode and ω_0 represents the angular frequency corres-

ponding to the frequency of oscillation and τ represents the transit time. Taking $E/E_{BT} = 0.9$ and η_1 corresponding to the efficiency at long wavelength, the difference in efficiencies $\eta_1 - \eta_2$ can be calculated. The result is given in Fig. 4 for the 304-B valve. It will be observed that there is fair agreement between theoretical and experimental values.

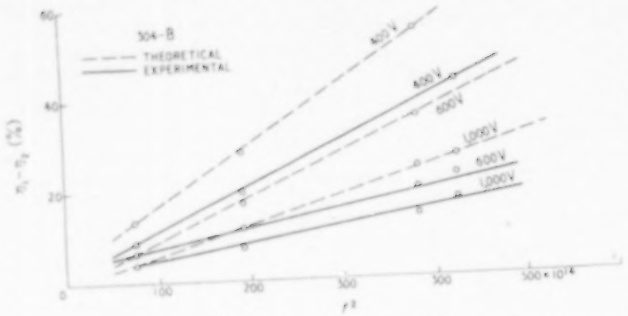

Fig. 4. Comparison of theoretical and experimental values of the fall of efficiency with f^2.

TABLE I

Transit angle corresponding to limiting frequency.

Valve	Anode voltage (V)	Transit angle (°)	Limiting frequency (Mc/s)
835	1,000	70	183.5
833-A	1,000	85	195.4
304-B	1,000	86	220.8
834	1,000	84	243.5
8012	1,000	>200	>600
316-A	400	130	661.2
955	200	110	669.1

TABLE II

Valve	λ_1 observed (cm)	λ_1 calculated (cm)	λ_2 observed (cm)	λ_2 calculated (cm)
835	600	539.9	171.4	174.4
833-A	600	657.7	176.5	219.2
304-B	448.9	505.9	145.9	168.7
834	397.9	467.9	129.3	152.5
8012	—	290.9	46.40	97
316-A	214.3	290.0	46.87	94.88
955	187.5	155.5	48	51.88

The wavelength λ_1 at which the efficiency drops by 10% of the maximum efficiency realizable can be calculated from Equs. (1) and (3). Similarly the wavelength λ_2 at which the efficiency is reduced to 10% of the maximum value can also be derived. For example, for valve 316-A,

$$\lambda_1 = \frac{2 \times 10^4 \times 0.29}{\sqrt{E_{BT}}} \text{ cm}$$

$$\lambda_2 = \frac{6.67 \times 10^3 \times 0.29}{\sqrt{E_{BT}}} \text{ cm}$$

The values of λ_1 and λ_2 both theoretical and experimental are given in Table II.

It will be observed that in almost all cases the calculated values of λ_1 and λ_2 are higher than the observed values. The difference, particularly in the cases of the 316-A and 8012 valves, is considerable. In the case of the 8012, the observed limiting frequency is nearly twice the calculated value. This is in agreement with the results obtained by Samuel[14] for double-ended operation.

Acknowledgment

The authors express their grateful thanks to the Head of the department for giving facilities to carry out the investigation.

APPENDIX

Transit time between cathode and grid :

e_a (Instantaneous anode voltage) $= E_{BT} - E_a \sin \omega t$

e_g (Instantaneous grid voltage) $= -E_{GB} + E_g \sin \omega t$

where E_a and E_g represent the amplitudes of the alternating components of the anode and grid voltages respectively, and E_{BT} and E_{GB} represent the steady supply voltages to anode and grid respectively.

Assume an electron leaving the cathode at time t_1 (i.e., when the grid just starts on its positive cycle) and arriving at the grid at an instant t_2. Then $\tau_{k_2} = t_2 - t_1$. At t_2

$$e_a = E_{BT} - E_a \sin \omega t_2$$
$$= E_{BT} - E_a \sin \omega (T/4 - t_g/2 + t_2 - t_1)$$
$$= E_{BT} - E_a \sin \omega (T/4 - t_g/2 + \tau_{k_2})$$

and $e_g = -E_{GB} + E_g \sin \omega (T/4 - t_g/2 + \tau_{k_2})$

The effective voltage on the grid at instant t_2 is given by

$$\left(e_g + \frac{e_a}{\mu}\right) = \left(-E_{GB} + \frac{E_{BT}}{\mu}\right) + \left(E_g - \frac{E_a}{\mu}\right)\left\{\sin \omega \left(\frac{T}{4} - \frac{t_g}{2} + \tau_{k_2}\right)\right\}$$

$$= \left(-E_{GB} + \frac{E_{BT}}{\mu}\right) + \left(E_g - \frac{E_a}{\mu}\right)\left\{\cos \left(\theta_{k_2} - \frac{\theta_g}{2}\right)\right\}$$

where, $\frac{2\pi}{T} \cdot \frac{t_g}{2} = \frac{\theta_g}{2}$ = Half the angle of grid current flow,

and $\frac{2\pi}{T} \cdot \tau_{k_2} = \theta_{k_2}$ = Transit angle.

Hence τ_{gs} given by Gavin[1] can be modified to

$$\tau_{gs} = \frac{3d_{gs}\sqrt{(m/2e)}}{\sqrt{\left[\left(-E_{GB}+\frac{E_{HT}}{\mu}\right)+\left(E_s-\frac{E_s}{\mu}\right)+\cos\left(\theta_{gs}-\frac{\theta_s}{2}\right)\right]}}$$

Transit time between grid and anode.

Let v be the velocity of an electron when it is at a distance x from the grid, at time t. Then the time of flight between grid and anode is given by the following expression:

$$\tau_{gs} = \int_{x=0}^{x=d_{gs}} \frac{dx}{v}$$

The potential at a point distant x from the grid at time t = potential at the grid plane at time t + potential gradient between plate and grid multiplied by x.

Therefore, the potential at x at time t is given by

$$V(x,t) = \left(e_s+\frac{e_s}{\mu}\right)+\left\{e_s-\left(e_s+\frac{e_s}{\mu}\right)\right\}\frac{x}{d_{gs}}$$

Writing,

$$e_s = -E_{GB}+E_s\cos\omega t$$

$$e_s = E_{HT}-E_s\cos\omega t$$

where $t = 0$ is the instant at which the electron arrives at the grid plane, which is the same as the instant at which the grid is at its positive peak, the expression for $V(x,t)$ reduces to

$$V(x,t) = \left[\left(\frac{E_{HT}}{\mu}-E_{GB}\right)+\left\{E_{HT}-\left(\frac{E_{HT}}{\mu}-E_{GB}\right)\right\}\frac{x}{d_{gs}}\right] + \left[\left(E_s-\frac{E_s}{\mu}\right)+\left\{-E_s-\left(E_s-\frac{E_s}{\mu}\right)\right\}\frac{x}{d_{gs}}\right].\cos\omega t$$

The velocity of the electron at x at time t is therefore given by

$$v(x,t) = \sqrt{\frac{2e}{m}}\cdot\sqrt{V(x,t)} = 5.93\times10^7\sqrt{V(x,t)}$$

At time $t = 0$, $\cos\omega t = 1$. In most cases τ_{gs} is of the order of 10^{-9} to 10^{-10} second. Therefore, in practical cases, as $\omega t = 2\pi\cdot t/T$, $\cos\omega t$ may be taken as unity up to 100 Mc/s. The expression for $V(x,t)$ then reduces to

$$V(x,t) = \left[\left(\frac{E_{HT}}{\mu}-E_{GB}\right)+\left(E_s-\frac{E_s}{\mu}\right)\right] + x\left[\frac{1}{d_{gs}}\left\{E_{HT}-\left(\frac{E_{HT}}{\mu}-E_{GB}\right)\right\}-\frac{1}{d_{gs}}\left\{E_s+\left(E_s-\frac{E_s}{\mu}\right)\right\}\right]$$

$$= A+Bx$$

$$\therefore \tau_{gs} = \int_{x=0}^{x=d_{gs}} \frac{dx}{5.93\times10^7\sqrt{(A+Bx)}} = \frac{2}{5.93\times10^7}\cdot\left\{\frac{\sqrt{(A+Bd_{gs})}-\sqrt{A}}{B}\right\}$$

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CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Secondary-Emission Tubes in Wideband Amplifiers

SIR,—With reference to the letter by N. F. Moody and G. J. R. McLusky on the above subject, published in your December issue, it may be of interest to note that the described method of obtaining push-pull output from anode and dynode of a secondary-emission valve was invented by E. V. Truefitt as early as 1937 when the TSE4 first became available. This invention is described in the specification of British Patent No. 508,038 which was published in 1939.

C. FALCONER CHAPTER

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London.

The Intrinsic Impedance of Space

SIR,—The fact that the technical literature mentions this impedance as being sometimes 377 ohms and at other times 30 ohms is misleading. I shall attempt to clear away this confusion.

The systems of units may differ both through the choice of the fundamental quantities and through the choice of the defining equations of the derived units so as to determine in a certain manner the value of the parasitic coefficients.

For instance, the area of a square which may be expressed in general through the formula, $s = al$, becomes specifically:

in the metric system:

$s = l^2$ square meters, where l is measured in meters,

in the Anglo-American system:

$s = l^2$ square inches, where l is measured in inches,

in the American circular system:

$s = \frac{4l^2}{\pi}$ circular inches, where l is measured in inches.

In the first two formulae the factor taken as equal to 1 is the parasitic coefficient of the formula giving the area of the square, whereas in the latter system the factor taken as equal to 1 is the coefficient of the formula giving the area of the circle.

The first choice is equivalent to taking the square of 1-m or 1-in side as the unit for areas, while the second one is equivalent to adopting the circle of 1-in diameter as the unit for areas.

Consequently, when the system of measuring units is changed, it must be expected, in general, that a number of units will have their magnitudes altered and some forms of the equations defining the corresponding quantities will be equally altered.

It is true that usually the systems are based on the

same forms of the defining equations, so that it is only a matter of changing the units when passing from one system to the other.

When, however, the problem of 'rationalizing' is raised in electricity and magnetism, the rationalized system introduced is different from the classical, non-rationalized system, not only as regards the units, but also in the forms of the defining equations.

The rationalizing may be effected in several ways, according to the formulae chosen, from which it is intended to derive the 4π factor.

In electricity and magnetism, it is noted sometimes that two quantities of different nature have the same dimension in a certain system of measuring units. For instance, in the systems using as fundamental quantities the length, the mass, the time and the intensity of the electric current, we encounter the following pairs of quantities having the same dimension :

(a) ϕ_D and q $[IT]$

(b) ϕ_B and m $[L^3MT^{-2}I^{-1}]$

(c) $Z_0 = \frac{\mathcal{E}}{H}$ and R $[L^2MT^{-2}I^{-2}]$

(d) J (intensity of magnetization) and B $[MT^{-2}I^{-1}]$

It would seem natural that these quantities should have units of the same magnitude in a particular system of units. But because of the parasitic coefficients of their defining formulae this is not the case. In some systems of units the two quantities of the same dimensions have identical units, whereas in other systems of units they have units of different magnitudes.

In the rationalized system we shall write :

$$\phi_D' = q$$

while in the non-rationalized system we write :

$$\phi_D = 4\pi q$$

While in the first system the unit of induction flow is the coulomb, in the second one it is the sterradian coulomb ; therefore, one coulomb $= 4\pi$ sterradian coulombs.

When we switch from the first system to the second one, the unit of quantity of electricity remains unchanged, while the unit of flow is changed, although these quantities always have the same dimension.

Therefore, the same flow will be expressed one time by ϕ_D and another time by ϕ_D', although the electric charge is all the time measured by q.

The same thing may be said of the units for ϕ_B and for m : the unit of magnetic flow in the m.k.s. system is the Weber in any of its two variants, whether rationalized or not. On the other hand, the magnetic mass defined by the relation :

$$\Sigma m' = \phi_B' \text{ in the rationalized system}$$

and $4\pi\Sigma m = \phi_B$ in the non-rationalized system will be expressed in webers in the rationalized system and in other units (newtons by millioersteds) in the non-rationalized system. This explains the different values assumed by Z_0, the intrinsic impedance of space, according to the system, rationalized or non-rationalized, in which it is expressed.

Z_0 is of the nature of an impedance and is defined by the equation :

$$Z_0 = \mathcal{E}/H$$

Since $\mathcal{E}$ remains unchanged when switching from the non-rationalized variant to the rationalized one, while H changes (in the non-rationalized system we have $\int H\,dl = 4\pi\Sigma I$ whereas in the rationalized system $\int H\,dl = \Sigma I$), it follows that Z_0 will have different units and consequently different values in each of the two variants.

In his Editorial of September 1949, Prof. G. W. O. Howe, reasoning on the case of a line consisting of two parallel flat strips,[1] shows that the intrinsic impedance $\mathcal{E}/H$ written in the rationalized system is expressed by :

$$Z = V/I$$

which is measured in ohms. On the basis of this practical observation we may draw the conclusion that Z_0 will be expressed in ohms only in the rationalized system (377 ohms), while in the non-rationalized system, where Z_0 has 30 as its numerical value, the units will be different (volts by millioersted and meter) :

$$\frac{\text{volts}}{\text{millioersted} \times \text{metre}} = 4\pi \text{ ohms.}$$

Thus, the intrinsic impedance of 30 volts by millioersted and meter, converted to ohms gives the same value of 377 ohms. Consequently, it may not be spoken of anything else than 377 ohms, *it being an error to say that the intrinsic impedance of space is* 30 *ohms*. If we change the definition of Z_0, stating that $Z_0 = \frac{\mathcal{E}}{H}$ in the rationalized system and $Z_0 = 4\pi\frac{\mathcal{E}}{H}$ in the non-rationalized system then the result in non-rationalized units is expressed equally in ohms, but its value is 377 and not 30 ohms.

T. TANASESCU

The Polytechnic Institute of Bucharest.

MONSIEUR,—Dans son numero de septembre dernier, votre Revue a publié sous le titre " The intrinsic impedance of space " et sous les initiales G.W.O.H., une critique de ma note du 18 février dernier à l'Académie des Sciences de Paris, dans laquelle le nom de M. Louis de Broglie a été accolé au mien, alors que, suivant l'usage français, son intervention s'est bornée à soumettre cette note au jugement de l' Académie. Monsieur de Broglie ne saurait donc encourir aucune responsabilité du fait de son contenu.

Je suis certain que vous voudrez bien faire connaître à vos lecteurs dans le plus bref délai, que le savant éminent qu'est Monsieur de Broglie ne devait pas être mis en cause dans cette discussion.

Je vous signale que l'erreur typographique concernant la valeur 477 ohms au lieu de 377 ohms a été signalée dans le numero de mai 1949 de la *Revue générale de l' Electricité*, page 168, note 1.

E. BRYLINSKI

Asnieres, Paris.

Detuned Resonant Circuits.

SIR,—The article under this title, by H. Elger, in the November 1949 issue of *Wireless Engineer* (p. 360), seems to contain a rather serious error in the determination of transient response. The author states that at high frequencies the phase of the exciting voltage at $t = 0$ can be assumed equal to zero ; this is remarkable, as surely the effect of the initial phase is the same in whatever frequency band the circuit is used. The initial phase has a negligible effect when the applied signal has the same frequency as resonance (i.e., $\omega = \omega_0$) so that equation (2c) is correct, but for all other frequencies of applied signal the effect is appreciable, so that equations (2a) and (2b) do not give an adequate statement of the response. This matter is brought out clearly in standard texts such as Guillemin[1] and in an article by myself.[2] If the suddenly-applied signal is proportional to $\sin(\omega t + \phi)$, and E is the steady-state peak output, then when ω is considerably different from ω_0, the tran-

[1] Guillemin, E. A., "Communication Networks," Vol. 1, p. 95 onwards (John Wiley, 1931).

[2] Tucker, D. G., "The Transient Response of a Tuned Circuit," *Electronic Engng*, 1946, Vol. 18, p. 379-381.

sient output e is given for the two extreme cases of initial phase, thus—

$$\left.\begin{aligned} \phi = 0, \quad & \frac{e}{E} \approx \cos \omega t - e^{-\omega_0 t/2Q} \cos \omega_0 t \\ \phi = \frac{\pi}{2}, \quad & \frac{e}{E} \approx \sin \omega t - \frac{\omega_0}{\omega} e^{-\omega_0 t/2Q} \sin \omega_0 t \end{aligned}\right\}$$

which should replace Mr. Elger's equation (2a), and the switching-off transients are

$$\left.\begin{aligned} \phi = 0, \quad & \frac{e}{E} = e^{-\omega_0 t/2Q} \cos \omega_0 t \\ \phi = \frac{\pi}{2}, \quad & \frac{e}{E} = \frac{\omega_0}{\omega} e^{-\omega_0 t/2Q} \sin \omega_0 t \end{aligned}\right\}$$

replacing equation (2b).

It can be seen that the purely transient component of the output can be affected in the ratio $\frac{\omega_0}{\omega}$ according to initial phase. For small detuning this may be an inappreciable effect, but Mr. Elger is certainly not justified in ignoring it when the detuning is 'excessive' (see last paragraph on p. 362). In actual fact, peak amplitudes when $\omega \ll \omega_0$ may run to ten or a hundred times the steady-state amplitude, and not just to twice, as Mr. Elger states. This is illustrated by oscillograms in my article. Moreover, in practice it is necessary to consider the response to signals outside the normal acceptance band, because although their steady-state effect may be negligible, the transient they produce may cause serious interference.

Post Office Research Station. D. G. TUCKER

SIR,—The mathematical statement of Mr. D. G. Tucker is quite correct and can easily be derived from my equation (1). In the publication[3] of Fraenckel it is shown by numerical examples that the peak voltage (or current) may materially exceed the value $2E$. My statement that the phase ϕ may be assumed to be equal to zero at high frequencies is, of course, not dependent on the frequency itself but on the practical use made of the different frequency ranges. It is evidently not necessary to explain the differences involved in problems relating to, say, power transmission at mains frequency and radar technique.

Mr. Tucker seems to have overlooked my statements preceding the equations (2a) and (3a) where I have expressly stated the conditions $\Delta\omega \ll \omega_0$ and $Q \gg 1$. At high frequency, 'excessive detuning' means any detuning beyond the bandwidth limits $Qv = 0.5$; i.e., a detuning exceeding 5% at $Q = 10$. In order to enable the reader to deal with other similar problems outside my theory I have indicated the equation (1).

My theory covers transients at least at $Qv = 4$ (see Fig. 4 and the well-known range of the normal universal resonance curve) $Q \geqslant 10$ and $\Delta\omega \ll \omega_0$. Equation (3a) in its present form yields $e_{max} = 2E$ and, if developed with $\phi = 90°$, $e_{max} = 1 + \omega_0/\omega$. If we now take the above extreme values applied to *any* frequency, we thus obtain $v_{max} = 25\%$ and the absolute maximum values $e = 2E$ and $e = 2.33E$ respectively and thus my 'serious error' can amount at most to nearly 17%; this means the extreme ends of my curve b in Fig. 4 may rise by 1.5 mm. Actually, the error will be much less since I have assumed the extreme and impossible case that the first peak occurs at $t = 0$ and thus coincides with the e/E axis of Fig. 3.

As to the last sentence of Mr. Tucker, he agrees with me that the response to signals (interference) outside the normal acceptance band should be considered and I had, therefore, shown the response within a range 8 times wider than the normal bandwidth. In practice it is evidently not necessary to calculate the response to interference far off the bandwith with any greater degree of accuracy if it is ever considered necessary to make numerical calculations at all. It is much more important to show that a clear and qualitative result is universally applicable (at the higher frequencies) and to give a distinct general conception of the transients in such cases.

I am, however, indebted to Mr. Tucker for drawing my attention to his publication and that of Guillemin which I had not seen until now due to particularly unfavourable circumstances at the time in question.

Stockholm, Sweden. H. ELGER

A Problem

A reader asks if anyone can suggest, or give a reference to a published description of, a method of finding the conditions under which

$$\int_{c-j\infty}^{c+j\infty} f(\lambda) e^{\lambda t} \, d\lambda$$

is monotonic in t and yet has the fastest rate of rise.—[ED.]

Phase-Shift Oscillator

SIR,—I was very interested to see W. C. Vaughan's analysis in the December *Wireless Engineer* which, although extremely ingenious, fails to take into account the internal impedance of the driving source.

The effect of this internal impedance can seldom be neglected, and after all, one of the chief advantages of the phase-shift oscillator is the precision with which it can be designed and made. Your readers may, therefore, be interested in an article of mine in the January 1950 issue of *Electronic Engineering* which demonstrates a comprehensive and flexible method of analysis by simple matrix algebra, and gives design formulæ which are free from approximations.

The matrix method is comprehensive because it provides the voltage ratio and input impedance of a four-terminal network simultaneously, and requires no special conditions concerning the similarity of transmission sections as it includes all valves and coupling arrangements.

The other advantage of flexibility enables a critical analysis to be made of an existing design, as it is comparatively easy to calculate the effect of changes in network elements.

Staines, Middlesex. W. R. HINTON

SIR,—Referring to W. C. Vaughan's article in the December 1949 *Wireless Engineer*, I feel that one general conclusion which might have been drawn from Tables I and II is that the networks containing shunt reactances may be used more profitably at high frequencies than those containing shunt resistances, while these latter are preferable at low frequencies. This is clear from the expressions for frequency since a greater CR product may be used in the former case than in the latter for a given frequency. This applies particularly to 3-mesh networks.

It is also suggested that the anode-load should be much less than the network input impedance—which might usefully have been tabulated—though no mention is made of a possibility of avoiding phase-shift in the valve altogether. In every case where shunt reactances are used the first resistance may be formed from the anode load in parallel with the valve's internal resistance. The first capacitance may then be formed, at least partly, by the anode-cathode capacitance and the last by the capacitance between grid and cathode. This is very useful

at high frequencies; e.g., for a 3-mesh network, if $R = 10^4$ MΩ, $C = 10$ pF and $n = 1$, $f = 3.9$ Mc/s.

Finally it is not obvious to me why the output waveform should be inherently better than with conventional LC circuits. This may be true at very low frequencies, but is surely not generally true.

Gt. Malvern, Worcs. W. P. N. COURT

Non-linear Effects in Rectifier Modulators.

SIR,—V. Belevitch gave a most interesting analysis of the nonlinear operation of a ring modulator in *Wireless Engineer*, May 1949, Vol. 26, p. 177, and stated also that the results applied equally to the bridge (otherwise known as shunt or Cowan) modulator with suitable adjustment, such as 6 db in the level of the products, etc. This analysis was based on a sinusoidal carrier and on the conception of the ideal rectifier, i.e., one with zero forward resistance and infinite backward resistance, changing over at zero voltage across it. Now this basis is very useful for analysing the performance of modulators in the linear condition, where the signal has no influence on the switching of the rectifier. However, we very much question its utility in the non-linear case, and would like to suggest that analytical results approaching those obtained in practice can be expected only by a more realistic approach.

No real rectifier switches from low to high resistance at the instant when the carrier (or bias) voltage changes polarity. There is always a transition interval, where non-linearity occurs not so much because the signal may exceed the carrier voltage and thus determine the time of 'switching' (as in Dr. Belevitch's ideal case), but rather because the rectifier resistance varies steeply

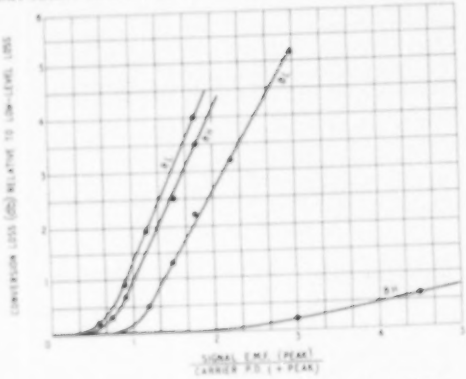

Fig. 1. *Overload characteristics of ring and bridge modulators.*

with voltage, and intermodulation and overloading occur. It is thus better to consider as an 'ideal' case for analysis one in which a square-wave carrier is used, so that in the absence of the signal the rectifiers switch from forward to backward resistance instantly. Whether the rectifiers are ideal or not has now less bearing on the result, which can consequently accord better with measurements. The mathematical analysis is no more complicated than in the sinusoidal case.

This conception of square-wave switching is approximated fairly closely by a practical case where a sinusoidal carrier e.m.f. is applied through a high resistance. We shall find that in such a case we obtain results very different from those given by Dr. Belevitch.

It is also inconvenient to consider rectifiers with zero forward resistance, since the carrier voltage can be developed across them only if the carrier source resistance (hereafter referred to as c.s.r.) is zero. Since, as we shall see, the question of c.s.r. is vital in the consideration of the bridge modulator, we should restrict our analysis to finite forward resistance.

In the ring modulator overload occurs when the signal voltage equals the carrier voltage across one of the rectifiers which is nominally in the blocking condition.

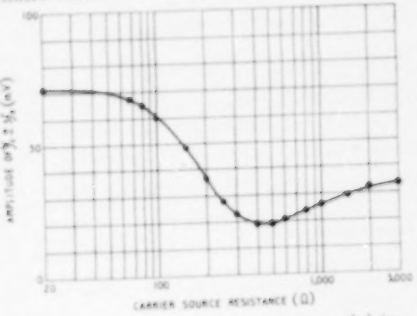

Fig. 2. *Harmonic distortion in bridge modulator.*

This does actually give an overload point corresponding very roughly to that shown by Dr. Belevitch. But in the conventional bridge modulator failure of the blocking condition in this way causes overload only in so far as the c.s.r., which is then thrown across the signal circuit, is low enough to prevent efficient modulation. Thus with a high c.s.r. overload cannot occur due to such a cause, and will only occur when the signal and carrier voltages across the forward resistance of each rectifier become equal—then one rectifier of each series pair becomes blocking and modulation fails. This occurs only at very high signal e.m.fs. The c.s.r. now limits the resistance to which the blocking due to overload can rise, and consequently a very high c.s.r. gives a more rapid overload than one less high. We thus find that there is an optimum c.s.r. which gives the highest overload point (and, in practice, the smallest nonlinear distortion below overload), and this has a value of the order of the circuit resistance (i.e., signal source and load in parallel).

Fig. 1 shows measured overload characteristics of ring (R) and bridge (B) modulators with low (L) and high (H) carrier source resistances. The terms 'low' and 'high' are relative to the forward resistance of the rectifiers which were copper-oxide type KG1; the carrier p.d. was 0.5 V peak for the ring and 1.0 V positive peak for the bridge (i.e., 0.5 V per rectifier in both cases), and the frequencies were 11 kc/s for the carrier and 500 c/s for the signal. The circuit resistance was 3,000 ohms, and the high c.s.r. was 1,000 ohms.

Fig. 2 shows measured harmonic distortion of the type $f_s = 3f_s$, in a bridge modulator, as a function of c.s.r. Here germanium rectifiers (CG1C) were used, with 2-V carrier (positive peak), 6-V signal e.m.f. and $R = 500$ ohms. It can be seen that the experimental results confirm the theoretical discussion.

A further point regarding bridge modulators is that the carrier voltage is greater across the backward than across the forward resistance unless steps are taken to equalize it by means of another rectifier or another modulator, connected in opposite polarity, in parallel with the carrier; this then becomes part of the c.s.r.

Post Office Research Station. D. G. TUCKER E. JEYNES

NEW BOOKS

Radio Aerials

By E. B. Moullin, M.A., Sc.D. Pp. 514 + xi. Oxford University Press, Amen House, Warwick Sq., London, E.C.4. Price 50s.

This is one of a series of monographs intended to provide accounts of recent advances in the scientific field. It deals only with a limited class of radio aerials, but does so in great detail.

The book is divided into two sections, one theoretical and the other experimental. In the former, after establishing Maxwell's equations and the Lorentz retarded potential functions the formulæ are used to calculate the fields due to current filaments, the effect of flat sheet and V reflectors and problems relating to cylinders immersed in electromagnetic fields. Arrays of radiating elements are then considered, the power gain calculated, and methods of suppressing the side lobes discussed. A short section is devoted to the isolated aerial, where attempts at a rigorous solution are mentioned, but not discussed in detail.

The remainder of the book is devoted to experimental procedure and measurements of aerial performance, mainly those made on the type of aerial described in the first section. Results for V reflectors are given in great detail, and many practical design considerations discussed.

The treatment is mainly mathemat cal; many of the problems concerned are hypothetical, and are included to illustrate the analytical methods. The solutions are later used to good effect, however, in obtaining approximate solutions of cases which cannot be solved rigorously, or where the rigorous solution would be too complicated to help in the understanding of the problem.

Throughout the book the author gives physical interpretations of the methods and results, which will appeal to the reader who likes to understand the significance of the mathematics as he goes along. This useful feature contributes greatly to the value and interest of the book.

In the calculation of radiation resistance, the author often refers to the value of the field at the surface of wires; inasmuch as no field tangential to the surface of a perfectly conducting wire can exist this statement needs qualification. If the postulated current distribution results in a tangential component of field, it means that it can only be sustained by using a series of generators distributed along the wire; the power supplied by these generators is then equal to the radiated power and is calculated by the method used by the author. If only one generator is supplied, the current distribution will adjust itself until the tangential component of field is zero along the wire.

This is, however, a minor criticism of a valuable addition to the literature on radio aerials. It is recommended, not only as a reference book on the types of aerial described, but also for the clear and logical development of the general theory.

H. P.

List of Preferred Valves

Pp. 28. The Scientific Instrument Manufacturers' Association of Great Britain Ltd., 17, Princes Gate, London, S.W.7. Price 2s. 6d.

This list of valves comprises types which are recommended by the Association for use in electronic instruments in order to limit the number of types used. The valves are chiefly ones with the IO and B7G bases. Abbreviated characteristics and base connections are included.

A Symposium on Electronics in Research a Industry

Edited by A. G. Peacock. Pp. 199 + xiii. Chapman & Hall Ltd., 37, Essex, St., London, W.C.2. Price 16s.

This book, which is published in collaboration with the Instrument Manufacturers' Association of Great Britain Ltd., comprises papers read at the first Electronics Symposium. The papers cover Electronics in Computing, Frequency Measurement, The Measurement of Small Displacements by Electrical and Electronic Methods, Measurement of Ionizing Radiations, High Vacuum Gauges, The Radiosonde and its Manufacture, Some Industrial Applications of Ultrasonics, Metal Detection in Industry, Sound and Vibration Measurement, Electronics in Spectroscopy, and Some Developments in Picture Telegraphy.

Electrical Transmission of Power and Signals

By Edward W. Kinibark. Pp. 461. John Wiley & Sons, Inc., New York, and Chapman & Hall Ltd., 37, Essex St., London, W.C.2. Price 48s.

This book, which is of American origin, covers r.f. transmission lines and waveguides, as well as power-frequency transmission.

Photoelectricity and its Application

By V. K. Zworykin, E.E., Ph.D., and E. G. Ramberg, Ph.D. Pp. 494 + xii. John Wiley & Sons, Inc., New York, and Chapman & Hall Ltd., 37, Essex St., London, W.C.2. Price 45s.

Theory and Desig n of Electron Beams

By J. R. Pierce. Pp. 197 + xiii. D. Van Nostrand Co. Inc., 250, Fourth Avenue, New York, U.S.A. Price $3.50.

Lateral Deviation of Radio Waves Reflected in the Ionosphere

By W. Ross, M.A., A.M.I.E.E. Radio Research Special Report No. 19. Pp. 32 + iv. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 9d. (U.S.A. 25 cents).

This report gives the results of investigations during the years 1938-1947 into the deviations from the great-circle plane of waves reflected from the ionosphere.

Television Explained (3rd Edition)

By W. E. Miller, M.A.(Cantab), M.Brit.I.R.E. Pp. 112 with 59 diagrams and 17 photographs. The Trader Publishing Co. Ltd., Dorset House, Stamford St., London, S.E.1. Price 5s.

Communication Circuits (3rd Edition)

By Lawrence A. Ware, E.E., Ph.D., and Henry R. Reed, M.S., E.E., Ph.D. Pp. 403 + x. John Wiley & Sons, Inc., New York, and Chapman & Hall Ltd., 37, Essex St., London, W.C.2. Price 40s.

F.B.I. REGISTER

The 1949/50 F.B.I. Register has over 1,000 pages and includes nearly 6,000 firms under 5,000 cross-reference headings. The sections include addresses, brands and trademarks, products and services. There are reference facilities in French and Spanish.

Published jointly by Kelly's Directories Ltd., and Iliffe & Sons Ltd. Home orders should be addressed to Kelly's Directories Ltd., 196, Strand, London, W.C.2, and overseas orders to Iliffe & Sons Ltd., Dorset House, Stamford St., London, S.E.1. The price is 42s.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.

AERIALS AND AERIAL SYSTEMS

625 413.—Aerial for aircraft wherein an insulated section of the aircraft skin has a metal film sprayed directly thereon.

Joshua Marney. Application date July 10th, 1947.

626 210.—A broadcast transmitting aerial employing a vertical radiator surrounded at its lower end by a concentric cage of subsidiary radiators, the latter having a radius of between $\lambda/100$ and $\lambda/50$.

Marconi's Wireless Telegraph Company Ltd., and H. Cafferata. Application date April 15th, 1946.

626 311.—An aerial system having a primary radiator and a reflector having a concave outer portion of parabolic form, designed so that the reflector has a ring focus.

Western Electric Company Inc. Convention date (U.S.A.) July 21st, 1945.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

626 236.—A pulse-radar system wherein the receiver embodies means for producing an oscillation at p.r. frequency (or related integrally thereto) which is combined with the received signals, the system being responsive only when the received pulse coincides in a particular manner with the local oscillation.

Marconi's Wireless Telegraph Company Ltd. (assignees of Thomas Treadwell Eaton). Convention date (U.S.A.) September 13th, 1941.

626 323.—A receiver and indicator for a beacon radiating a beam varying in frequency according to azimuth direction, and having two selective means differently variable in attenuation with frequency, the outputs from the two means being compared.

Standard Telephones and Cables Ltd. (assignees of A. Frum). Convention date (U.S.A.) June 11th, 1945.

626 610.—F.M. radio altimeter in which a voltage dependent on the speed of the modulator capacitor drive motor provides a compensating function to render this system independent of modulation frequency.

Bernard Anselm Sharpe. Application date May 3rd, 1945.

626 709.—V.H.F. high pulse-power generators using several waveguides of different propagation times and feeding into a common waveguide.

Compagnie General De Telegraphie Sans Fil. Convention date (France) March 4th, 1944.

626 870.—Receiving apparatus for beacons wherein the phase of signals from spaced transmitters is compared, consisting in using aerials having different patterns which are employed successively to avoid erroneous phase indications.

William Joseph O'Brien. Convention date (U.S.A.) August 27th, 1945.

626 956.—A radio beacon for aircraft distance indication emitting primary signals in successive directions and having distinguishing characteristics, the aircraft emitting signals having a different characteristic, the beacon retransmitting response signals to the aircraft.

Standard Telephones and Cables Limited (assignees of H. G. Busignies). Convention date (U.S.A.) August 11th, 1945.

RECEIVING CIRCUITS AND APPARATUS

(See also under Television)

626 088.—Automatic gain or frequency control in which the valve to which the control voltage is applied includes a phase-shift network to introduce a virtual capacitance into the smoothing circuit to vary the time constant thereof.

Philips Lamps Limited. Convention date (Netherlands) June 1st, 1943.

626 142.—Thermionic valve apparatus embodying independent screening chassis each of which carries a part of a valveholder, the other part of which is carried by another chassis.

Philips Lamps Ltd. Convention date (Netherlands) September 14th, 1943.

626 154.—Phase or frequency-modulated signal receiver operating on the super-regenerative principle and designed primarily to secure low distortion.

Hazeltine Corporation (assignees of B. D. Loughlin). Convention date (U.S.A.) March 19th, 1946.

626 221.—Signal-seeking receiver having motor tuning which stops on tuning to a usable signal, wherein control of the motor is made insensitive to noise by a narrow-band selector.

Colonial Radio Corporation. Convention date (U.S.A.) September 16th, 1944.

626 420.—Phase or frequency-modulation detector especially for converting such pulses to amplitude-modulated pulses, using two rectifiers and a cathode-follower stage.

Bendix Aviation Corporation. Convention date (U.S.A.) August 23rd, 1946.

626 739.—Frequency discriminator in which tuned circuits feed opposed rectifiers and also two further opposite rectifiers, the two sets of rectifiers feeding a metering circuit to indicate or monitor the modulation.

Marconi's Wireless Telegraph Company Ltd. and I. S. Forbes. Application date February 6th, 1947.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

626 715.—Apparatus for television film transmission using two film projectors for alternative use and adjustable means for transmitting from either projector to the television camera.

H. G. De France. Convention dates (France) October 11th, 1941, August 12th, 1942.

SUBSIDIARY APPARATUS AND MATERIALS

626 129.—Permeability-tuned circuits in which coils and cores are mounted for relative movement and coupled by an auxiliary coil arranged so that the coupling coefficient varies inversely with frequency.

Marconi's Wireless Telegraph Co. Ltd. (assignees H. Ruben). Convention date (U.S.A.) May 12th, 1945.

626 620.—The dielectric of a capacitor for medium frequencies consists of an aliphatic ester of boric acid having prescribed characteristics.

The British Thomson-Houston Co. Ltd. Convention date (U.S.A.) September 1st, 1944.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

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ACOUSTICS AND AUDIO FREQUENCIES

534.321.9 269

Ultrasonics: A Brief Survey.—J. H. Jupe. (*Electronic Engng.*, Nov. 1949, Vol. 21, No. 261, pp. 422–423. Bibliography, pp. 423–424.)

534.78 270

Speech Communication under Conditions of Deafness or Loud Noise.—W. G. Radley. (*Proc. Instn elect. Engrs*, Part I, Nov. 1949, Vol. 96, No. 102, pp. 312–313.) Discussion on 2690 of 1948.

534.78 271

The Sonograph: Elements and Principles.—J. Dreyfus-Graf. (*Schweiz. Arch. angew. Wiss. Tech.*, Dec. 1948, Vol. 14, No. 12, pp. 353–362. In French.) Description of an instrument which transforms each sound into a group of electrical pulses and then into mechanical motions characterizing certain elements of the sound. It can transform speech into writing and may have wide application in linguistic investigations.

534.851 : 621.395.813 272

Measuring Turntable Speed Fluctuations.—E. W. Berth-Jones. (*Wireless World*, Dec. 1949, Vol. 55, No. 12, pp. 471–474.) Fluctuations of the order of 0.01% must be detected. An a.f. oscillator is used which generates a tone of known frequency, constant within less than 0.01%. This tone is fed through the recording channel to the cutter head, which is mounted in position on the recording lathe to be tested. A pickup is mounted so that it will track the groove cut by the recording head, at a distance of a few inches behind it. The output of the pickup is amplified and matched in level with a second output obtained from the recording channel. If these outputs are adjusted to be in phase initially, they can be combined to have zero resultant at constant turntable speed; small variations in this speed cause the resultant to have an amplitude dependent on the change in speed. The method can be adapted to magnetic as well as disk recording. Sensitivity can be adjusted by varying the oscillator frequency.

621.395.625 273

The Development of Mobile Recording Technique.—M. J. L. Pulling. (*B.B.C. Quart.*, Oct. 1949, Vol. 4, No. 3, pp. 179–192.) Description of various types of apparatus used by the British Broadcasting Corporation since 1936, notably the Type-C equipment which operated from two 6-V accumulators in series and could be carried in a private car, and the Midget Disk Recorder which weighs 35 lb complete. Magnetic recorders have not so far been used much; they have marked advantages for recording sporting events where important incidents occur infrequently and unexpectedly. Problems of organization and possible future improvements are discussed. See also 11 of 1949.

621.395.625.2 : 621.396.97 274

Reproduction from Disks and Records for Broadcasting.—J. W. Godfrey. (*B.B.C. Quart.*, Oct. 1949, Vol. 4, No. 3, pp. 170–175.) Variations in recording standards and in reproducing devices, particularly playback needles, make it difficult to obtain consistency in quality and signal/noise ratio. A reasonable compromise is possible if light-weight pickups and a standardized sapphire-tipped reproducing stylus are used, and if selective equalization is used in reproducing circuits. The effects of variations in playback time, in the linear speed of the groove, in recording characteristic, and in needle and groove radius are considered separately.

681.85 : 621.317.616† 275

The Variable-Disk-Speed Method of Measuring the Frequency Characteristics of Pick-Ups.—P. R. Terry. (*B.B.C. Quart.*, Oct. 1949, Vol. 4, No. 3, pp. 176–178.) A test disk has a single band of recorded tone, and its speed of rotation is varied in a 4 : 1 range. A portion of the pickup frequency-response curve is thus obtained, which can easily be joined to similar portions at other frequencies. Although the response of a pickup cannot be fully defined in a simple manner, this method gives adequate information for practical purposes without requiring calibrated test disks.

AERIALS AND TRANSMISSION LINES

621.315.212 : 621.397.5 276

London-Birmingham Television Cable.—H. Stanesby & W. K. Weston. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 186–200.) Reprint. See 1279 and 1857 of 1949.

621.315.65 277

Notes on a Coaxial Line Bead.—D. W. Peterson. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1294.) Comment on 1582 of 1949 (Cornes).

621.392.26 278

Electromagnetic Waves in Metal Tubes.—K. Honerjäger. (*FIAT Review of German Science, 1939–1946. Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 31–44. In German.) A concise treatment of the propagation of E-, H-, and Lecher-type waves, which are grouped under the term "line waves" (Leitungswellen). The principal formulae for the propagation constants for waveguides of rectangular or circular cross-section and for concentric cables are summarized. The excitation of line waves by electric or magnetic dipoles, and their reflection, refraction and diffraction, are also considered. References to over 30 articles and reviews published in Germany are given.

621.392.26† : 621.3.09 279

The Propagation of Electromagnetic Waves in a Cylindrical Tube containing a Coaxial D.C. Discharge.—P. Rosen. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 865–877.) The boundary-value problem is solved. Ohm's law is assumed to hold for the a.c.; this has been verified theoretically for the case where the a.c. field is small compared with the d.c. field. "Curves which give the relationship between the complex propagation constant, γ, and the complex dielectric coefficient, K_c', have been computed for the solution in which the TEM mode of the coaxial line is approached as K_c' becomes infinite."

621.392.26† : 621.396.662 280

Corrections to the Attenuation Constants of Piston Attenuators.—J. Brown. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 491–495.) The modes existing in a circular waveguide with walls of finite conductivity are investigated and the effect of an oxide layer on the inner surface of the attenuator is considered. Expressions are derived for the corrections to the attenuation constant, which are of importance in the case of H modes when an accuracy within 1 part in 10^4 is desired.

621.396.67 : 621.315.14 281

Antennas and Open-Wire Lines: Part 1—Theory and Summary of Measurements.—R. King. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 832–850.) The apparent impedance of an aerial as a load on an open-wire transmission line may be determined from the theoretical impedance of the isolated aerial if the transmission line end-effects and the coupling between the line and the aerial are represented by an inductance in series and a capacitance in parallel with the aerial. The inductive effects are small, while the capacitive effects may be large, so that the general variation of the apparent aerial impedance, with various types of transmission line and methods of connection, may be derived by determining the effect of a lumped positive or negative capacitance in parallel with the isolated aerial. Curves are given showing the good agreement obtained between the experimental and theoretical results. Part 2, 399 below.

621.396.67 282

Aerials.—K. Fränz. (*FIAT Review of German Science 1939–1946. Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 65–89. In German.) Discussion of the properties of dipole, long-wire, and parabolic aerials, slot and dielectric radiators, and of various measurement methods, with references to 26 relevant papers.

621.396.67 283

The Influence of Conductor Size on the Properties of Helical Beam Antennas.—T. E. Tice & J. D. Kraus. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1296.) Results of measurements on three helices of identical construction, except for conductor diameter, indicate that the frequency range of the beam mode is only slightly affected by conductor diameter. See also 306 and 1860 of 1949 (Kraus).

621.396.67 : 538.566.2 284

The Magnetic Dipole in a Stratified Atmosphere.—G. Eckart. (*Onde élect.*, Oct. 1949, Vol. 29, No. 271, pp. 378–381.) The e.m. field is determined for a dipole in a free atmosphere whose dielectric constant is a linear function of the z coordinate. The case of such an atmosphere above a plane earth is also considered. The physical interpretation of the results is discussed.

621.396.67 : 551.510.535 285

Impedance Characteristics of Some Experimental Broad-Band Antennas for Vertical Incidence Ionosphere Sounding.—H. N. Cones. (*Bur. Stand. J. Res.*, July 1949, Vol. 43, No. 1, pp. 71–78.) Results of measurements of the modulus of input impedance of a number of non-resonant aerials over a continuous frequency range from 1 to 25 Mc/s are shown graphically and discussed. The aerials are compared with each other from the standpoint of uniformity of impedance over the frequency range. The use of multiple-wire construction to lower the average input impedance, to minimize impedance variations, and to increase radiation efficiency is considered.

621.396.67 : 629.135 286

Suppressed Aircraft Aerials.—G. E. Beck. (*Wireless World*, Dec. 1949, Vol. 55, No. 12, pp. 468–470.) Drag is reduced by using (*a*) the whole aircraft as an aerial, excitation being provided by a small coil in the root of the wing, (*b*) rod or loop aerials buried in a portion of the wing or fuselage, any drag-producing cavities being filled in with a woven-glass type of dielectric, (*c*) slot aerials. Each of these methods is briefly discussed. See also 1335 of 1947 (Booker) and 2457 of 1948 (Johnson).

621.396.671 287

Gain of Aerial Systems.—J. Brown. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 409–410.) When an aperture of given size is used as an end-fire radiator instead of as a broad-side radiator, only the normal increase in gain is realized which occurs when a radiator is situated at the surface of a perfect reflector. This increase is less than that predicted by Bell (3057 of 1949), because the field strength near the plane of the aperture is not zero, at any rate in the direction of the beam.

621.396.671 288

Input Impedance of Wide-Angle Conical Antennas Fed by a Coaxial Line.—C. H. Papas & R. King. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1269–1271.) Such impedances have been computed for several flare angles; certain auxiliary functions are shown graphically.

621.396.671 : 621.317.336 289

Antennas and Open-Wire Lines: Part 2—Measurements on Two-Wire Lines.—Tomiyasu. (*See* 399.)

621.396.677 290

Analysis of the Metal-Strip Delay Structure for Microwave Lenses.—S. B. Cohn. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, p. 1011.) Addendum to 2147 of 1949.

621.396.679.4 291

Open-Wire Line for F.M.—J. W. Ecklin. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 80–81.) Two 0-gauge wires, spaced 6-in. apart, were used for connecting a 10-kW transmitter to a high-gain aerial at the top of a 240-ft. tower. Matching was effected at each end of the line by means of a bazooka, which is essentially a 1:1 transformer for taking care of the balanced-to-unbalanced to ground conditions, together with a $\lambda/4$ coaxial impedance-matching section. This arrangement

is cheaper and more efficient than a comparable coaxial feeder and is substantially unaffected by the weather. Construction and installation details are discussed.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.3 292

"G" Curves in Tube Circuit Design.—K. A. Pullen. (*Tele-Tech*, July & Aug. 1949, Vol. 8, Nos. 7 & 8, pp. 34–36 & 33–35, 59.) Curves and abacs for the determination of dynamic operating characteristics of valve circuits directly and for evaluation of distortion. Applications to various amplifier and oscillator circuits are discussed.

621.3.016.352 293

A Generalization of Nyquist's Stability Criteria.—A. Varsonyi. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 863–867.) A new type of diagram is developed, from which the lower limits of the 'damping ratios' can be determined, and hence the degree of stability of systems characterized by ordinary linear differential equations. See also 294 below.

621.3.016.352 294

A Generalization of Nyquist's Stability Criteria.—S. J. Mason. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, p. 867.) Comment on 293 above. An alternative method of Q-determination is described in which the ordinary Nyquist diagram is used.

621.3.018.83† 295

R-Q Factor.—W. W. Harman. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1295.) The relation between the Q of an electrical resonator and its resonance or shunt resistance R can be written in the form $R = \zeta Q/\omega_0 C$, where $0 < \zeta < 1$. The RQ factor ζ is a measure of how effectively the electric field in the resonator is concentrated in the capacitance.

621.314.2.045 : 621.3.011.4 296

Winding Capacitance.—N. H. Crowhurst. (*Electronic Engng*, Nov. 1949, Vol. 21, No. 261, pp. 417–421, 431.) Discussion of winding capacitance for interleaved or random-wound multilayer a.f. transformers, chokes etc. Charts are given for determining distributed, interwinding or winding-to-screen capacitance. Practical examples are considered.

621.314.26 : 621.396.645.371 297

Frequency Changers and Amplifiers with Constant Gain.—D. G. Tucker. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1324–1327.) A method of applying negative feedback to a frequency changer by using a similar frequency changer in the feedback path is described. If the reduction of conversion gain due to the feedback is adjusted to 6 db, equal changes of ± 2 db in each frequency changer only alter the overall conversion gain by ± 0.1 db. A typical pentode circuit is described. Nonlinear distortion is reduced by the feedback. The application of the principle to stable-gain linear amplifiers is discussed; constancy of gain compares favourably with that of conventional feedback systems, but the conditions for linearity are more difficult to realize.

621.314.3† 298

Barkhausen Noise and Magnetic Amplifiers: Part 2 — Analysis of the Noise.—J. A. Krumhansl & R. T. Beyer. (*J. appl. Phys.*, June 1949, Vol. 20, No. 6, pp. 582–586.) The Barkhausen noise is calculated for ferromagnetic cores by a method similar to that used for analysing the shot effect. The signal/noise ratio of a typical amplifier circuit is calculated and compared with the measured value. Part 1 : 2730 of 1949.

621.314.3† 299

On the No-Load Characteristics of a Transductor.—K. Kühnert & M. Delattre. (*C. R. Acad. Sci., Paris*, 17th Oct. 1949, Vol. 229, No. 16, pp. 751–753.)

621.316.8 300

Fixed Resistors for use in Communication Equipment.—P. R. Coursey. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, p. 482.) Discussion on 2735 of 1949.

621.316.8 301

Resistors for Deposited-Circuit Techniques.—W. R. Conway. (*Electronic Engng*, Nov. 1949, Vol. 21, No. 261, pp. 403–408.) Charts are given from which the dimensions of a rectangular film resistor of given wattage can be determined in terms of the aspect ratio. The case of fractional electrodes is also considered.

621.316.86 302

Investigations on Carbon-Layer Resistors.—A. Schulze & D. Bender. (*Elektrotechnik, Berlin*, Oct. 1947, Vol. 1, No. 4, pp. 97–105.) A short general discussion of methods of production, properties, and methods of measurement, with experimental curves showing the dependence of the resistance on age, temperature, humidity, and loading.

621.316.86 303

Negative Temperature Coefficient Resistors.—(*Philips' tech. Commun., Aust.*, 1949, Nos. 2/3, pp. 35–39.) The resistors consist of mixed crystals of Fe_3O_4 and other spinels, such as $MgAl_2O_4$ and Zn_2TiO_4. Properties and applications are discussed.

621.318.572 : 621.396.1 304

Electronic Diversity Switching.—H. V. Griffiths & R. W. Bayliff. (*Wireless World*, Nov. & Dec. 1949, Vol. 55, Nos. 11 & 12, pp. 414–418 & 486–488.) The previous diversity switching system (1878 of 1949) required Type EF8 valves. Since these valves are no longer in production and no other type is available with a similar low ratio of screen to anode current, a modified dual-diversity switch using perferred types of valve was designed. This makes some simplification of the prototype triple-diversity system possible. Advantages over combined diversity systems are discussed.

621.392 305

Network Theorem.—E. R. Wigan. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, p. 409.) The theorem enunciated is derived very simply from the basic equations which define the properties of a quadripole. It states that the ratio between the output open-circuit voltage and the input voltage is identically equal to the ratio of the output short-circuit current to the input current, the direction of transmission of power through the network being reversed in the second case. An example of the application of the theorem is given.

621.392 306

A Note on Thévenin's Theorem.—(*Electrician*, 4th Nov. 1949, Vol. 143, No. 3725, pp. 1473–1474.) A simple proof, based on Kirchhoff's laws.

621.392 : 517.512.2 307

Application of Fourier Transforms to Variable-Frequency Circuit Analysis.—A. G. Clavier. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1287–1290.) The behaviour of passive circuits is studied for the case where f.m. is applied to the driving force. The output current or voltage is expressed in the form of a convolution integral, which can lead either to the expansion of Carson & Fry (464 of 1938) or preferably to that of van der Pol (2310 of 1946); the latter is expressed in terms of the values of the transfer im-

pedance or admittance for the instantaneous frequency, and its derivatives. Convergence conditions are discussed. The particular case of broad-band f.m. line discriminators, for which the convolution integral can be expressed in terms of known functions, is considered. See also 1800 of 1943.

621.392 : 621.385.3

308

The Application of Power Series to the Solution of Non-Linear Circuit Problems.—A. W. Gillies. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 453–475.) Discussion of circuit problems in which a triode is associated with any kind of linear network. Carson's power series solution for a triode circuit can be extended to cover such cases. See also 3825 of 1945 (Tucker) and 2740 of 1948 (Cartwright).

Part 1: Triode Circuit with Negative Feedback. The reduction of nonlinear distortion in an amplifier by means of negative feedback is discussed. Conditions of stability are derived. The power-series solution remains convergent even when the circuit is regenerative, failing completely only at the point of critical regeneration. When the circuit is unstable and an e.m.f. of nearly the natural frequency is applied, the solution represents an unstable oscillation, but when the frequency of the applied e.m.f. is not near the resonance frequency, the solution represents a stable forced oscillation if the amplitude is sufficiently large.

Part 2: The Free Oscillations of a Regenerative Triode Circuit. A single complex equation determines the frequency of oscillation and the amplitude of the fundamental; the complex amplitudes of the harmonics are expressed by power series. A particular circuit is considered in detail.

Part 3: The Forced Oscillations of a Regenerative Triode Circuit. The conditions for synchronization and for the suppression of the free oscillation are discussed, and also the asymmetry of the resonance curves. The analysis is extended to harmonic and subharmonic resonance, and the mechanism by which the various effects are produced is explained.

621.392 : 621.396.619.13

309

The Solution of Steady-State Problems in F.M.—B. Gold. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1264–1269.) A method is described for deriving a polynomial approximation to the admittance of a given network. The degree of the polynomial depends on the ratio of the width of the sideband spectrum to the bandwidth of the network. This polynomial approximation can be used with advantage when the sideband method is impracticable and the quasi-steady-state method invalid; it enables the output wave to be expressed in terms of the derivatives of the input wave.

If the range of approximation coincides with the part of the spectrum of the input wave that is not negligible, and if the approximating function is derived from a set of orthogonal polynomials, a practical method for solving steady-state f.m. problems can be established. The validity of this method is investigated and examples are included.

621.392.5

310

Delay Networks having Maximally Flat Frequency Characteristics.—W. E. Thomson. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 487–490.) A lumped-constant equivalent of a transmission line can be obtained in general in the form of a symmetrical lattice, in which the series and lattice arms are inverse and approximate respectively to the short-circuit and open-circuit impedances of half the line. One such set of approximations can be derived from the infinite ladder networks (Cauer's canonical form) equivalent to these impedances.

"These approximations produce all-pass constant impedance networks (dissipation being neglected) in which the delay is maximally flat in the sense that the first $2m-1$ derivatives of the delay with respect to frequency are zero at the origin; m is an integer expressing the order of the approximation."

621.392.52

311

A Generalized Formula for Recurrent Filters.—M. C. Pease. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1293.) A generalization of the formula given by Fano & Lawson (695 of 1948).

621.392.52

312

Impedance Transformations in Four-Element Band-Pass Filters.—R. O. Rowlands. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1337–1340.) By using a different type of basic section for constructing the filter, a similar transformation to that described by Belevitch (3818 of 1947) can be achieved with slightly greater economy, both in particular cases and in the general case.

621.392.6

313

Theory of 2n-Pole Networks.—R. Feldtkeller. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 91–102. In German.) General review, with references to 56 relevant German papers. A note is added by N. T. Ming with reference to unpublished work, from papers left by W. Cauer, on (*a*) the practical calculation of reactance, bridge, and recurrent quadripoles for given characteristics, (*b*) the design of filters with prescribed minimum damping curve in the blocking region, (*c*) the design of amplifiers with oscillatory-circuit coupling and with prescribed amplitude characteristics, (*d*) the calculation of h.f. amplifiers with relatively narrow pass-band.

621.395.645.37 : 621.395.44

314

Transmitting Amplifier for the K2 Carrier System.—H. C. Fleming. (*Bell Lab. Rec.*, Nov. 1949, Vol. 27, No. 11, pp. 391–393.) The line amplifiers for this system have thermistor control that keeps their output approximately constant. The amplifier at the transmitting terminal must therefore have a constant output, however many channels are in use. This is achieved by using the amplifier also as a 60-kc/s oscillator; its 60-kc/s output is varied so that the combined output, consisting of voice sidebands plus 60-c/s signal, remains nearly constant. The method is explained with the aid of circuit diagrams.

621.396.029.65†

315

Millimetre Waves: A General Survey.—A. W. Lines. (*T.R.E. J.*, July 1949, pp. 1–20.) Discussion of: (*a*) the behaviour of the klystron and the rising-sun magnetron at mm λ, (*b*) other possible types of mm-wave generator, (*c*) atmospheric absorption, (*d*) the hobbing technique for anode construction, (*e*) possible methods of increasing resonator size or resonator separation, (*f*) tunable reflex klystrons for λ 8–9 mm, (*g*) frequency instability, (*h*) techniques for constructing small waveguides, (*i*) techniques for propagation in larger waveguides, (*j*) optical methods of transmission, (*k*) reception technique, for which the crystal valve and reflex klystron local oscillator, with a balanced mixer, are essential, (*l*) methods of measurement.

621.396.611.1 + 621.392.52

316

Resonance Curves from Tables of Functions, and Some Simplifications in the Theory of Electrical Filters.—H. Nitz. (*Frequenz*, Aug. 1949, Vol. 3, No. 8, pp. 237–244.) Simple series and parallel oscillatory circuits are considered and expressions are derived for their apparent impedance. By suitable transformations

these expressions are put into a form which enables both branches of the resonance curve to be determined directly from tables of functions. Essential simplification of known formulae for filter characteristics results from similar transformations, including the transformation of hyperbolic functions into a form scarcely used hitherto. These transformations give an appreciably better insight into the functional behaviour of many characteristics and in conjunction with Hayashi's tables of functions reduce the time required for numerical computations. The methods are applied to T and Π filters of general form and to the simplification of certain quadripole formulae. The functions used are tabulated.

317

621.396.611.1

Frequency Contours for Microwave Oscillator with Resonant Load.—M. S. Wheeler. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1332–1336.) Discussion of a method of representing frequency and power relations in an oscillator coupled through a transmission line to a frequency-sensitive load. The method is applied to the case of a reactance valve coupled to a magnetron, and a solution is obtained, making certain simplifying assumptions. A more general method, not requiring these assumptions, is then developed and used to determine the limiting conditions for the appearance of frequency discontinuities in the characteristics.

318

621.396.611.4

Theory of Systems with Transmission-Line Coupling.—A. Weissfloch. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 103–119. In German.) Discussion with special reference to 4-pole and 6-pole devices for decimetre and centimetre waves. 35 relevant German papers are noted.

319

621.396.611.4

Cavity Resonators.—R. Müller. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 44–48. In German.) A short review, with references to 28 relevant German papers.

320

621.396.611.4

Theory of Cavity-Resonator Systems.—G. Goubau. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 120–126. In German.) Short discussion of the coupling theory developed by Dahlke, the cavity resonator being regarded as a quasi-stationary oscillatory circuit, and an outline of the $2n$-pole theory developed by Goubau.

321

621.396.611.4

On the Experimental Determination of the Resonance Resistance of E.M. Cavities.—F. Borgnis. (*Helv. phys. Acta*, 15th Oct. 1949, Vol. 22, No. 5, pp. 555–578. In German.) Various general methods available for the determination of resonance resistance, using microwave technique, are discussed and applied to cavity resonators, for which the e.m. field in certain regions, to which the resonance resistance is related, can be regarded as either homogeneous or rotationally symmetrical. The necessary theoretical relationships are derived; experimental investigations for λ 14 cm on resonators of different shapes confirm the suitability and the accuracy of the methods considered.

322

621.396.611.4 : 621.384.611.2†

Quarter-Wavelength Coaxial-Line Resonators for Betatron-Started Synchrotrons.—Goward, Wilkins, Holmes & Watson. (*See* 4305.)

323

621.396.615

Design of Nonlinear Sine-Wave Oscillators by the Moving-Axis Method.—J. Abelé. (*Ann. Phys., Paris*, Nov.–Dec. 1948, Vol. 3, pp. 655–679.) The amplitude of an oscillation represented by a differential equation of the second or higher order is defined as the locus of the maxima of a family of oscillations. This definition is applied to a linear oscillator, and a geometrical construction for determining the amplitude is described. This is extended to a certain class of nonlinear oscillators which are capable of steady-state oscillations that are rigorously sinusoidal; such oscillators include those with amplitude control. See also 2525 of 1946.

324

621.396.615

Phase-Shift Oscillator.—W. C. Vaughan. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 391–399.) An examination of the mechanism of three- and four-mesh phase-shifting networks. The condition considered is that of steady oscillation, assuming that (*a*) oscillations can be initiated, (*b*) the waveform is sinusoidal, (*c*) the load resistance is substantially less than the input impedance of the network. From simple derivations of the values of voltage and current for each component in the network, three tables are compiled: (*a*) formulae for the frequency at which 180° phase reversal occurs, and the corresponding relations which define the voltage loss in a four-mesh network as each shunt and series impedance is varied in turn, (*b*) corresponding formulae for a three-mesh network, (*c*) the voltage and current formulae for a uniform three-mesh phase-advancing network. See also 1298 of 1948 (Dawe).

325

621.396.615

A Variable Phase-Shift Frequency-Modulated Oscillator.—O. E. De Lange. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1328–1330.) The theory of operation is discussed for an oscillator consisting of a broad-band amplifier whose output is fed back to the input through a phase-shifting circuit. The instantaneous frequency is controlled by the phase shift. The frequency deviation is directly proportional to the instantaneous amplitude of the modulating signal and substantially independent of the modulation frequency. A practical 65-Mc/s circuit is described.

326

621.396.615

The Reactance-Tube Oscillator.—H. Chang & V. C. Rideout. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1330–1331.) A single-valve combination of a capacitive or inductive reactance-valve circuit and an oscillator circuit. The capacitive type is similar to the Hartley oscillator and the inductive to the Colpitts oscillator. A linear frequency/grid-voltage relation and constant output amplitude can be obtained over a range of more than 5% around centre frequencies in the range 1–4 Mc/s.

327

621.396.615.17

The Blocking Oscillator as a Variable-Frequency Source.—L. Fleming. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1293.) A blocking oscillator can be used instead of the positive-bias multivibrator described by Bertram (1884 of 1948) to transform variations of a d.c. voltage into variations in the frequency of a sub-carrier.

328

621.396.615.17 : 621.317.755

The Integration Method of Linearizing Exponential Waveforms.—A. W. Keen. (*J. Brit. Instn Radio Engrs*, Nov. 1949, Vol. 9, No. 11, pp. 414–423.) The exponential response of the simple CR integrator to a step voltage differs from linearity by an error voltage proportional to the integral of the exponential output. Three methods of nearly cancelling this error voltage by means of an additional CR integrating circuit are given. Each

corrected network has an LR equivalent and may be arranged for voltage or current excitation. Practical arrangements are discussed for (*a*) linear sawtooth voltage generation, (*b*) waveform linearization in time-base voltage amplifiers, (*c*) linear sawtooth current generation, as in television reception timebases. In case (*c*), split deflecting coils are used.

621.396.645 329

Valve Amplifiers for Decimetre Waves.—W. Bürck. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 1–11. In German.) The special valve and circuit problems for dm-λ apparatus are discussed and an illustrated description is given of a 3-stage amplifier for λ 59 cm. This uses disk valves, Type LD12, and tubular tank circuits, and has a power amplification of about 10 per stage. Another amplifier, using two push-pull Type-LV4 pentodes and with a power amplification of 3–4 per stage and a bandwidth of about 8 Mc/s, is mentioned. Development of valves of the ceramic-disk type enabled amplifiers to be constructed with a tuning range of 48–53 cm, a bandwidth of about 1 Mc/s and overall power amplification of about 400 for 2 stages.

621.396.645 330

Various Types of Amplifier.—K. Schönhammer. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 12–19. In German.) Brief discussion, with references to 64 papers published in Germany, of investigations on d.c. amplification, broad-band amplification, gain limits, linearity, feedback, and on amplifiers for voice-frequency and carrier-frequency telephony, for the transmission of speech and music, and for measurement purposes.

621.396.645 331

A New Wide-Band Amplifier Circuit.—R. Aschen. (*T.S.F. pour Tous*, Nov. 1949, Vol. 25, No. 253, pp. 349–352.) The principles and construction are described of an aperiodic amplifier with a remarkably linear gain curve from 25 c/s to above 10 Mc/s. The two valves used are Types EF42 and EL41, which give a gain of 20. The valves have a common load resistor and strong negative feedback from one anode to the other. Followed by a sensitive detector using a double triode, Type ECC40, and giving full-scale deflection on a 100-μA meter for an applied voltage of 250 mV, the amplifier constitutes an aperiodic monitor with many practical applications.

621.396.645 332

Distributed Amplification.—T. Sárkány. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1294.) Comment on 3375 of 1948 (Ginzton et al.).

621.396.645 333

Amplification of E.M. Waves by Interaction between Electron Beams under the Influence of Crossed Electric and Magnetic Fields.—R. Warnecke, O. Doehler & W. Kleen. (*C. R. Acad. Sci., Paris*, 10th Oct. 1949, Vol. 229, No. 15, pp. 709–710.) Increased amplification can be obtained by an arrangement such that the electrons move in a direction normal to the directions of the crossed fields. A plane structure is here considered in which there are two narrow parallel beams between parallel electrodes; a formula is derived for the complex propagation constant. The expression for the imaginary part of this constant may explain certain phenomena associated with magnetrons.

621.396.645 : 537.311.33 : 621.315.59 334

Some Circuit Aspects of the Transistor.—R. M. Ryder & R. J. Kircher. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 367–400.) An analysis of the type-A transistor as an active quadripole is given and the equivalent circuits for single-stage and cascade amplifiers are discussed in relation to gain, stability and termination. Gains of 15–20 db per stage can be realized and examination of the dependence of the frequency response on collector voltage and point spacing shows that useful gain is possible at frequencies up to 10 Mc/s. Large-signal operation is discussed with reference to the power output and the distortion in amplifiers and oscillators; power transistors giving outputs of 200–600 mW are described.

Over the frequency range 20–20 000 c/s the noise power per unit bandwidth is proportional to (frequency)$^{-1.1}$ and appears to be a mixture of smooth and impulse noise. The noise factor at 1 000 c/s is about 60 db.

621.396.645.029.3 335

High-Quality Amplifier: New Version: Parts 2-4.—D. T. N. Williamson. (*Wireless World*, Oct.–Dec. 1949, Vol. 55, Nos. 10–12, pp. 305–309, 423–427 & 477–479.) Design details of tone controls, auxiliary gramophone circuits, and a complete tone-compensation and filter unit, with the circuit of a receiver for use in districts where the spacing between the carrier frequencies of the principal local transmitters is of the order of 200 kc/s. Part 1: 3101 of 1949. See also 70 of January (Sarser & Sprinkle).

621.396.645.36 : 621.385 336

Secondary Emission Tubes in Wideband Amplifiers.—N. F. Moody & G. J. R. McLusky. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 410–411.) Push-pull output can be obtained from a single valve, by fitting appropriate loads in series with both dynode and anode. The application of this principle to the Type-EF150 valves of an amplifier with distributed amplification is briefly considered. See also 3375 of 1948 (Ginzton et al.).

621.396.662 : 621.392.26† 337

Magnetically Controlled Wave-Guide Attenuators.—T. Miller. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 878–883.) Discussion of experiments on the power loss in a waveguide filled with various iron powders, and the variations due to an external magnetic field applied either parallel or perpendicular to the magnetic component of the e.m. wave in the guide. Effects are observed which may be due to ferromagnetic resonance. A theoretical formula for the power loss of low-conductivity powders is developed and compared with experimental results.

621.396.69 338

Circuit Components for V.H.F.—H. Meinke. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 48–64. In German.) Reflection-free elements of uniform transmission lines and waveguides, transformers, connections between conductors of different types (such as balanced and unbalanced lines or concentric and parallel twin conductors), reflection-free terminating impedors, voltage dividers, chokes and resonance circuits are considered. References are given to 68 German publications.

621.397.645 339

Cathode Neutralization of Video Amplifiers.—J. M. Miller, Jr. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1345.) Addendum to 3393 of 1949.

621.318.2 340

Permanent Magnets. [Book Review]—F. G. Spreadbury. Publishers: Pitman & Sons, Ltd, London, 280 pp., 35s. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, p. 411.) "A well-written, up-to-date and comprehensive work covering fundamental theory, materials, circuit design,

applications, measurements etc. The treatment of the subjects is concise and appropriately mathematical or descriptive.

621.318.4 + 621.314.2 341

Einführung in die Theorie der Spulen und Übertrager mit Eisenblechkernen (Introduction to the Theory of Coils and Transformers with Laminated-Iron Cores). [Book Review.]—R. Feldtkeller. Publishers: S. Hirzel Verlag, Stuttgart, 2nd edn. Part 1, 190 pp., 10.50 DM. Part 2, 130 pp., 8 DM. Part 3, 65 pp., 4 DM. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 411–412.) The second edition, revised and extended, of one of a series of monographs dealing with telegraphy and telephony. It treats the subject both mathematically and from the practical point of view for the telecommunications engineer. The three parts deal with (*a*) coils, (*b*) transformers, (*c*) design data. "The book is . . . a valuable addition to the available information on this subject."

621.392 : 51 342

The Mathematics of Circuit Analysis. [Book Review.]—Guillemin. (*See* 389.)

GENERAL PHYSICS

530.12 : 531.18 : 537.122 343

Special Relativity and the Electron.—W. W. Harman. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1308–1314.) An elementary paper, intended to introduce the subject "in a manner attractive to the engineer". See also 82 of January (Kübler.)

535.412 344

Study of Interference Fringes in the Neighbourhood of Caustics.—E. Durand. (*Ann. Phys., Paris*, Nov./Dec. 1948, Vol. 3, pp. 621–636.) The interference fringes near the caustic of a concave spherical mirror inclined at 45° to an incident beam of parallel rays is considered. The intensities in the system of fringes are calculated by geometrical optics and the interferences by Huyghens' principle; results are verified experimentally.

535.42 345

On the General Laws of Diffraction. Critical Review.—G. Toraldo di Francia. (*Rev. d'Optique*, Nov. 1949, Vol. 28, No. 11, pp. 597–611.) Difficulties encountered in the practical application of various classic theories are discussed. Different expressions for Huyghens' principle are considered; that of Luneberg appears to be most suitable for present-day applications. Luneberg has chosen a single-layer and double-layer distribution on the surface of integration which differs from that of Helmholtz and Kirchhoff. The principle of inverse interference, which is related to Huyghens' principle, is used to explain (*a*) the existence of evanescent waves, (*b*) transmission and total reflection phenomena, (*c*) diffraction of non-planar incident waves by grids, (*d*) a thermodynamic conception of diffraction.

535.8 346

On the Corrector Plates of Schmidt Cameras.—E. H. Linfoot & E. Wolf. (*J. opt. Soc. Amer.*, Sept. 1949, Vol. 39, No. 9, pp. 752–756.) Discussion of the design of the aspherical surface of a Schmidt corrector so as to obtain optimum performance, in an agreed sense, over the field taken as a whole.

535.8 347

On the Optics of the Schmidt Camera.—E. H. Linfoot. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 3, pp. 279–297.) "A fifth-order aberration function is derived for the ordinary Schmidt camera. Colour correction is also considered and a proof is given of J. G. Baker's formulae for the first three coefficients in the aspheric plate profile of a system with minimized axial colour-spreads (*Proc. Amer. phil. Soc.*, 1940, Vol. 82, pp. 323–338). Aberration functions are used to design a modified system in which the monochromatic aberrations are balanced over a finite field, and the performance of this system is compared with that of the ordinary Schmidt camera by means of spot diagrams."

537.213 : 517.564.4 348

The Application of Non-Integral Legendre Functions to Potential Problems.—Hall. (*See* 384.)

537.311.4 : 621.315.59 349

On the Theory of the A.C. Impedance of a Contact Rectifier.—J. Bardeen. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 428–434.) An extension of Dilworth's d.c. analysis (2769 of 1948) to the a.c. case, leading to the usual equivalent circuit (R and C in parallel, with R, in series) which was discussed by Spenke in an earlier comprehensive analysis (532 of 1942). R and C may depend upon the d.c. flowing and also on frequency, because of the effect of ionic drift.

537.311.5 + 537.312.6 350

Distribution of Temperature and Current in Cylindrical Bodies.—C. F. Muckenhoupt. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 939–942.) These distributions depend not only on skin effect but also on the variation of resistivity with temperature. Since the resistivity change is 3–4% for every 10°C rise in temperature for metals, and greater for insulators, its effect can be appreciable. A complete solution for the steady state is given.

537.525 : 621.396.61 351

Relaxation Oscillations in Discharge Tubes: Application to the Study of [discharge] **Ignition.**—J. Moussiegt. (*Ann. Phys., Paris*, Sept./Oct. 1949, Vol. 4, pp. 593–670.) A detailed investigation is described of the phenomena associated with intermittent discharges in a commercial neon tube with twisted stranded electrodes lying along the axis, their free ends being about 3 mm apart. The critical capacitance value for the production of relaxation oscillations and the conditions during the relaxation cycle are particularly considered. Previous theories of the intermittent discharge are discussed and a theory is developed which involves a time constant for the ignition period, assuming an exponential law. This theory is confirmed for the discharge tube with axial electrodes. For other types, with different electrode structures, the theory certainly holds good in some cases. Values of the ignition time constant, determined from measurements of current maxima, are given.

538.5 352

The Induction of Electric Current in Non-Uniform Thin Sheets and Shells.—A. T. Price. (*Quart. J. Mech. appl. Math.*, Sept. 1949, Vol. 2, Part 3, pp. 283–310.) General equations are obtained for the induction of electric currents, in any thin-sheet distribution of conducting material, by periodic or aperiodic fields. Methods of solving these equations are considered, with special reference to plane sheets and spherical shells.

538.541 353

On the Theory of the Eddy-Current Anomaly.—R. Feldtkeller. (*Frequenz*, Aug. 1949, Vol. 3, No. 8, pp. 229–237.) Any explanation of this anomaly for ordinary transformer sheet must take account of the effect of variation of the initial permeability and of the doubling of the field strength from the surface to the middle of the material.

538.56.029.64 : 531.61 354

Torque and Angular Momentum of Centimetre Electromagnetic Waves.—N. Carrara. (*Nature, Lond.*, 19th Nov. 1949, Vol. 164, No. 4177, pp. 882–884.) When a plane circularly polarized e.m. wave is absorbed by a screen at right angles to the direction of propagation, it exerts a mechanical torque S/ω per unit surface, where

S is the Poynting vector and $\omega/2\pi$ is the frequency. Beth (2451 of 1936) measured this torque for light waves; a method of measuring it for centimetre waves is here discussed. Results are in qualitative agreement with theory; the fact that circularly polarized waves possess angular momentum is established.

538.3 355

The Fundamentals of Electromagnetism. [Book Review]—E. G. Cullwick. Publishers: Cambridge University Press, 2nd edn, 327 pp., 18s. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 383–384.) For comment on first edition see 4764 of 1939 and 1260 of 1940.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72+523.854+523.53 : 621.396.822 356

Radio Astronomy.—J. S. Hey. (*Nature, Lond.*, 12th Nov. 1949, Vol. 164, No. 4176, pp. 815–817.) Report on a British Association meeting at which recent progress was reviewed. Summaries are included of: (*a*) a paper by J. S. Hey describing investigations undertaken by the Army Operational Research Group on r.f. emission from sunspots and solar flares, (*b*) a paper by F. G. Smith on recent investigations at the Cavendish Laboratory, Cambridge, on galactic r.f. radiation, (*c*) a paper by A. C. B. Lovell on recent research in meteor astronomy at the University of Manchester by means of radio reflection or radar methods.

523.75 357

The Emission of Radiation from Flares.—R. G. Giovanelli. (*Mon. Not. R. astr. Soc.*, 1949, Vol. 109, No. 3, pp. 337–342.) "An extension of recent theories of the emission and absorption of radiation indicates that the temperature in a typical flare is $< 2 \times 10^5$ deg. K and the electron concentration is $< 10^{13}$ per cc."

551.510.535 : 621.3.087.4 358

A Single-Band 0–20-Mc/s Ionosphere Recorder embodying some New Techniques.—Wadley. (*See* 392.)

551.510.535 : 621.396.11 359

Thermal Expansion of Ionospheric Layer and Temporary Morning Disappearance of Radio Signals.—Banerjee & Singh. (*See* 440.)

551.576 : 621.396.9 360

Application of Radar Equipment to Storm Location in South East Asia.—Lutkin & Chisholm. (*See* 362.)

551.594.6 : 621.396.821 361

Atmospherics.—H. Siedentopf. (*FIAT Review of German Science, 1939–1946. Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 167–171. In German.) Short discussion of different types, their relation to weather conditions, and their seasonal variations.

LOCATION AND AIDS TO NAVIGATION

621.396.9 : 551.576 362

Application of Radar Equipment to Storm Location in South East Asia.—F. E. Lutkin & J. Chisholm. (*T.R.E. J.*, July 1949, pp. 21–30.) The American 40-kW 3-cm radar Type AN/APS-15 was fitted in a vehicle, with a 30-in. scanner mounted on the roof, at each of 7 Royal Air Force meteorological stations. Plan presentation of clouds within 50 miles was obtained, though response was reduced at extreme range when there was rain near the observing station. On Singapore Island, an A.M.E.S. Type 13 radar was installed; this 500-kW 10-cm set has a narrow beam in the vertical plane which sweeps over 25° in elevation. Maximum range was 150 miles; cloud responses and permanent echoes could be distinguished. The general nature of storm clouds in this region and of their radar responses is discussed. There was good correlation between cloud reports by pilots or from local meteorological stations and those given by the equipments; the radar information both increased the forecaster's knowledge of atmospheric processes and was of immediate practical value for short-term forecasts for aviation.

621.396.93 363

Direction Finding.—P. v. Handel. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 173–183. In German.) The term 'direction finding' is used in the wider sense to include bearing and distance measurements with h.f. waves. Small-base arrangements, in which the base length is small compared with λ, are first considered, then large-base systems and finally methods of distance measurement. References are given to 35 relevant German publications.

621.396.93 364

Some Relations between Speed of Indication, Bandwidth, and Signal-to-Random-Noise Ratio in Radio Navigation and Direction Finding.—H. Busignies & M. Dishal. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 228–242.) Reprint. See 2232 of 1949.

621.396.933 : 621.396.619.16 365

Pulse-Multiplex System for Distance-Measuring Equipment (DME).—C. J. Hirsch. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1236–1242.) Distance from a ground transponder beacon is determined by measuring the time interval between sending an interrogator pulse and receiving the reply. Traffic among several beacons with overlapping service areas can be handled by frequency and pulse-pair coding arrangements so that each signal consists of a pulse pair of distinctive spacing, of the order of 10–25 μs. Circuits are described which recognize signals having only one such spacing. The equipment described is for 52 channels, but the same method could be used for 100 channels. See also 102 of January (Burgmann).

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 366

McLeod-Type Alloy-Filled Vacuum Gauge.—J. Groszkowski. (*Nature, Lond.*, 19th Nov. 1949, Vol. 164, No. 4177, pp. 886–887.) Certain disadvantages associated with the Hg-filled McLeod gauge can be avoided if Hg is replaced by an easily fused alloy, such as that comprising 27% Pb, 13% Sn, 50% Bi and 10% Cd; the effects of this change on gauge design are discussed.

535.37 367

Dielectric Changes in Phosphors containing more than One Activator.—G. F. J. Garlick & A. F. Gibson. (*Proc. phys. Soc.*, 1st Nov. 1949, Vol. 62, No. 359A, pp. 731–736.) Measurements of these dielectric changes are discussed. Assuming that the changes are due to electron trapping, information regarding the nature of the traps and their apparent association with luminescence centres is derived.

535.37 : 535.61-15 368

The Rise in Brightness of Infra-Red-Sensitive Phosphors.—R. C. Herman & C. F. Meyer. (*J. opt. Soc. Amer.*, Sept. 1949, Vol. 39, No. 9, pp. 729–731.) "A phenomenological theory of infra-red-sensitive phosphors is given which assumes the existence of two types of luminescent centers in order to account for the so-called 'inertia' effects in the rise of brightness. The equations describing the electron transfer processes have been integrated and show under certain conditions a very rapid initial rise in brightness followed by a relatively slow attainment of the maximum. The effect of varying the intensity of the stimulating infra-red radiation is discussed."

535.61-15 : 535.215 369

The Temperature Variation of the Long-Wave Limit of Infra-Red Photoconductivity in Lead Sulphide and Similar Substances.—T. S. Moss. (*Proc. phys. Soc.*, 1st Nov. 1949, Vol. 62, No. 359B, pp. 741–748.)

538.221 370

Ferro-Magnetism.—(*Elect. Times*, 17th Nov. 1949, Vol. 116, No. 3028, pp. 687–688.) Discussions at an I.E.E. symposium.

538.221 371

Ferroxcube.—(*Philips tech. Commun., Aust.*, 1949, Nos. 2/3, pp. 28–34.) Manufacture, properties and applications of two varieties of low-loss ferrites.

538.221 372

Magnetic Characteristics of an Oriented 50-Percent Nickel-Iron Alloy.—J. H. Crede & J. P. Martin. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 966–971.) Single-crystal magnetic properties have been closely approached in a polycrystalline 50%-Ni 50%-Fe alloy by the development of a favourable grain orientation. Elimination of the first and third steps of the normal magnetization process produced a hysteresis loop of nearly rectangular shape.

538.221 373

Magnetic Viscosity in Mn-Zn Ferrite.—R. Street & J. C. Woolley. (*Proc. phys. Soc.*, 1st Nov. 1949, Vol. 62, No. 359A, pp. 743–745.)

546.431.82 : 536.48 374

Symmetry Changes in Barium Titanate at Low Temperatures and their Relation to its Ferroelectric Properties.—H. F. Kay & P. Vousden. (*Phil. Mag.*, Oct. 1949, Vol. 40, No. 309, pp. 1019–1040.) The optical changes in $BaTiO_3$ are described; they can be completely explained if the crystal symmetry changes from tetragonal to orthorhombic at −5°C, and then to rhombohedral at −90°C. These changes have been confirmed by X-ray investigations; they are due to successive spontaneous polarizations along the [100], [110] and [111] cube directions. The relation of polarization to the cell structure is discussed; the simple Lorentz equation does not apply if the Ti/O interaction energy is large. The three transitions are explained on the basis of this interaction; difficulties of electrical dipole co-operative effects in $BaTiO_3$ are considered.

549.514.51 375

Increase in Q-Value and Reduction of Aging of Quartz Crystal Blanks.—A. C. Prichard, M. A. A. Druesne & D. G. McCaa. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, p. 1011.) The quartz blank is annealed by heating it almost to the inversion temperature of quartz, or 500°C, and cooling it down extremely slowly. Values of Q thus obtained are at least double those of untreated quartz blanks, and variations in frequency and Q are minute. These improvements appear to be permanent. A more detailed report is being prepared.

549.514.51 376

Salvaging Electrically Twinned Quartz.—J. L. Rycroft & L. A. Thomas. (*Electronic Engng*, Nov. 1949, Vol. 21, No. 261, pp. 410–415. Correction, *ibid.*, Dec. 1949, Vol. 21, No. 262, p. 477.) For another account see 112 of 1949 (Wooster, Wooster, Rycroft & Thomas).

621.315.59 377

Editorial Note regarding Semiconductors.—(*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 335–343.) Surveys briefly the atomic physics of impurity semiconductors, with special reference to Si and Ge and the phenomena of rectification and amplification in semiconducting devices.

621.315.59 378

Theory of Transient Phenomena in the Transport of Holes in an Excess Semiconductor.—C. Herring. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 401–427.) "An analysis is given of the transient behaviour of the density of holes n_x in an excess semiconductor as a function of time t and of position x with respect to the electrode from which they are being injected. When the geometry is one-dimensional, an exact solution for the function $n_x(x,t)$ can be constructed, provided certain simplifying assumptions are fulfilled, of which the most important are that there be no appreciable trapping of holes or electrons and that diffusion be negligible. An attempt is made to estimate the range of conditions over which the neglect of diffusion will be justified. A few applications of the theory to possible experiments are discussed."

621.315.59 : 537.311.33 : 621.396.645 379

The Theory of *p-n* Junctions in Semiconductors and *p-n* Junction Transistors.—W. Shockley. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 435–489.) "In a single crystal of semiconductor the impurity concentration may vary from *p*-type to *n*-type producing a mechanically continuous rectifying junction. The theory of potential distribution and rectification for *p-n* junctions is developed with emphasis on germanium. The currents across the junction are carried by the diffusion of holes in *n*-type material and electrons in *p*-type material, resulting in an admittance for a simple case varying as $(1 + i\omega\tau_p)^{\frac{1}{2}}$ where τ_p is the lifetime of a hole in the *n*-region. Contact potentials across *p-n* junctions, involving no current, may develop when hole or electron injection occurs. The principles and theory of a *p-n-p* transistor are described."

621.315.59 : 537.311.33 : 621.396.645 380

Hole Injection in Germanium — Quantitative Studies and Filamentary Transistors.—W. Shockley, G. L. Pearson & J. R. Haynes. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 344–366.) Holes injected by an emitter point into thin single-crystal filaments of germanium can be detected by collector points. From studies of transient phenomena the drift velocity and lifetimes (as long as 140 μs) can be directly observed and the mobility measured. Hole concentrations and hole currents are measured in terms of the modulation of the conductivity produced by their presence. Filamentary transistors utilizing this modulation of conductivity are described.

621.315.611.011.5 : 548.0 381

Polarizability and Dielectric Constant of Ionic Crystals.—B. Szigeti. (*Trans. Faraday Soc.*, Feb. 1949, Vol. 45, No. 314, pp. 155–166.) The polarizability α of a crystal is connected with its natural frequencies. In static fields or for very long external waves it depends on the shape of the material, but for short waves it depends only upon whether the wave is longitudinal or transverse. The Clausius-Mosotti formula connecting α with the dielectric constant χ is $(\epsilon-1)/(\epsilon+2) = 4\pi\alpha/3$; it holds for a sphere in very long waves. The Drude formula $\epsilon - 1 = 4\pi\alpha$ is valid for short transverse waves. For short longitudinal waves, $4\pi\alpha = (\epsilon-1)/\epsilon\pi$.

669.177 382

Electrolytic Iron.—C. Tschäppät. (*Schweiz. Arch. angew. Wiss. Tech.*, Aug. 1949, Vol. 15, No. 8, pp. 225–242. In French.) The production of thin sheets of electrolytic iron has hitherto been a matter of great difficulty. An account is given of investigations, extending over the last 20 years, with the object of producing thin sheets directly by electrolysis, thus eliminating melting and rolling processes. The principles of the process evolved are explained and the mechanical and electrical properties of the sheets produced are described. Such sheets of thickness 0.05–0.3 mm appear to be ideal for many electrotechnical applications as well as for the production of small mechanical parts.

383

621.774.7

Treatise on Powder Metallurgy: Vol. 1 — Technology of Metal Powders and their Products. [Book Review]—C. G. Goetzel. Publishers: Interscience Publishers, London, 778 pp., 56/-. (*Metal Ind., Lond.*, 9th Dec. 1949, Vol. 75, No. 24, pp. 495–496.) The first of 3 volumes planned "to organize the mass of present-day knowledge on powder metallurgy in a standard manner". A short introductory chapter maps out the whole field; succeeding chapters fill in the details. ". . . this work can be unreservedly recommended to all who have any interest in powder metallurgy."

MATHEMATICS

384

517.564.4 : 537.213

The Application of Non-Integral Legendre Functions to Potential Problems.—R. S. Hall. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 925–931.) The numerical evaluation of various potentials requires knowledge of zero-order Legendre functions of real but non-integral degree. Tables and curves are given for the zeros of these functions and for certain integrals, together with a number of associated approximate formulae. Illustrative examples involving conducting cones, spheres and rings are included.

385

517.942 : 537.566

On Weber's Function.—C. G. Darwin. (*Quart. J. Mech. appl. Math.*, Sept. 1949, Vol. 2, Part 3, pp. 311–320.) Discussion of the behaviour of real solutions of the differential equation

$$(d^2u/dx^2) + (\tfrac{1}{4}x^2 - a)u = 0$$

which occurs in connection with e.m. wave propagation in the ionosphere and in vibration problems. Convergent power series are given for the two solutions, and also associated asymptotic series for large values of x and for large values of a. Tables of the real solutions are being prepared.

386

681.142

An Electronic Differential Analyzer.—A. B. Macnee. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1315–1324.) The analyser described can be used to solve ordinary differential equations, both linear and nonlinear, of orders up to and including the sixth. The coefficients may be constant or variable. The analyser has a high speed of operation and is extremely flexible. Questions of periodicity and stability and of the continuity of solutions can be investigated, as well as problems in which final rather than initial values are specified. New types of electronic function generator and of electronic multiplier are used. Accuracy is within 1–5%; solutions can be repeated within 0.002–0.1%. Sources of error are analysed.

387

681.142

The Bell Relay Computer.—F. L. Alt. (*Instruments*, Oct. 1948, Vol. 21, No. 10, pp. 912–913.) A brief illustrated discussion. This computer is a great deal slower than the ENIAC, but much more flexible; it can be instructed to choose between alternative courses of action. See also I.R.E. paper by G. R. Stibitz entitled "Counting Computers", of which a summary was noted in 1793 of 1949.

388

51 : 621.39

Compléments de Mathématiques à l'Usage des Ingénieurs de l'Électrotechnique et des Télécommunications (Mathematics for Electrical and Telecommunications Engineers.) [Book Review]—A. Angot. Publishers: Éditions de la Revue d'Optique, Paris, 1949, 660 pp. (*Nature, Lond.*, 12th Nov. 1949, Vol. 164, No. 4176, pp. 809–810.) In the reviewer's opinion the book provides "a common language" for the pure mathematician and the practical man. It shows both the relevant applications of theory and the theoretical basis of familiar practical results. Many subjects are covered. Fundamental groundwork is included for each, with full discussion of appropriate numerical examples. A sufficient basis for the reader's further progress by himself is thus provided. The book is "recommended both as a textbook and especially as a reference book".

389

51 : 621.392

The Mathematics of Circuit Analysis. [Book Review]—E. A. Guillemin. Publishers: J. Wiley & Sons, New York, 1949, 575 pp., $7.50. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1304: *J. Franklin Inst.*, Oct. 1949, Vol. 248, No. 4, pp. 356–357.) "It is not a mathematical textbook on the principles of circuit analysis, but rather . . . a presentation, within the compass of a single volume, of material appropriate to provide for the student fundamental mathematical equipment for the understanding of the theory underlying advances in the subject of circuit analysis. Applications of the theory are reserved for later publications . . . Each chapter furnishes a well-rounded treatment of its subject."

390

517.948

Integralgleichungen (Integral Equations). [Book Review]—G. Hamel. Publishers: Springer-Verlag, Berlin-Göttingen-Heidelberg, 2nd edn 1949, 166 pp., 13.60 DM. (*Phys. Blätter*, 1949, Vol. 5, No. 10, pp. 481–482.) Theory and applications. Based on a course of external lectures at the Technical College, Berlin.

MEASUREMENTS AND TEST GEAR

391

531.761

Direct Reading Timer and Clock.—A. E. Wolfe, Jr. & F. G. Steele. (*Radio & Televis. News, Radio-Electronic Engng Supplement*, Nov. 1949, Vol. 13, No. 5, pp. 3–5, 28.) Design and construction details of an electronic clock for measuring intervals from 0.01 sec to 24 hr. For another account see *Electronics*, Dec. 1949, Vol. 22, No. 12, pp. 75–77.

392

621.3.087.4 : 551.510.535

A Single-Band 0 - 20-Mc/s Ionosphere Recorder embodying some New Techniques.—T. L. Wadley. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 483–486.) Description of a recorder developed in South Africa. See also 229 of 1948 (F. J. Hewitt, J. Hewitt & J. Wadley).

393

621.317.3

Measurement of [electrical] **Constants of Materials.**—C. Schmelzer. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 268–278. In German.) A review of various methods for measurement of dielectric constant, loss factor, h.f. resistance etc., including methods suitable for u.h.f. References to 32 relevant German publications are given.

394

621.317.3 : 621.396.11

Wave-Propagation Measurements.—Beckmann. (*See* 441.)

395

621.317.324†

Measurement of Field Distribution.—A. Stenzel. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 264–268. In German.) A short account of the "test-body" method used first by Müller for loss-free cavity resonators (1379 of 1940) and extended by Goubau to the general linear 2-pole (1870 of 1944).

396

621.317.335.3† : 621.315.615

The Measurement of Dielectric Constants of Liquids by a Frequency Deviation Method.—W. L. G. Gent. (*Trans. Faraday Soc.*, Aug. 1949, Vol. 45, No. 320, pp. 758–

759.) A method which allows a continuous check to be made on the standard capacitance by calibration against a crystal oscillator.

621.317.335.3† + 621.317.374 : 621.396.611.4 **397**

Measurement of the Dielectric Constant and Loss of Solids and Liquids by a Cavity Perturbation Method.—G. Birnbaum & J. Franceau. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 817–818.) The changes Δf, ΔQ in the resonance frequency and Q of a cavity resonator when a small cylindrical sample of a solid is inserted are measured by a method in which the resonance curve, together with a pair of calibrated variable frequency markers, are displayed on a c.r.o. screen. Formulae due to Bethe & Schwinger relate Δf and ΔQ to the complex dielectric constant of the solid. A block diagram of the equipment is given. Typical results are tabulated and compared with those of Bleaney, Loubser & Penrose (3187 of 1947). The method described extends the usefulness of Sproull & Linder's method (2240 of 1946) by its sensitive technique for measuring small frequency differences.

621.317.335 **398**

Impedance Measurements.—A. Weissfloch. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 234–241. In German.) A general description of methods suitable for use in the dm-λ and cm-λ regions. The methods mainly use resonance effects or the properties of transmission lines. References to 14 relevant German papers are given.

621.317.336 : 621.396.671 **399**

Antennas and Open-Wire Lines: Part 2 — Measurements on Two-Wire Lines.—K. Tomiyasu. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 892–896.) Difficulties involved in such measurements are discussed. These include the problem of balance. The impedance of aerials is shown to depend on the nature of the driving structures. Measured impedances agree well with theoretical values. Part 1: 281 above.

621.317.372 **400**

Microwave Q Measurements in the Presence of Series Losses.—L. Malter & G. R. Brewer. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 918–925.) Appreciable errors can result from neglecting the losses in coupling devices, which appear in equivalent circuits in the form of series resistance. Formulae are derived and curves are plotted for determining Q and the circuit efficiency, given the s.w.r. at resonance and far from resonance, and knowing whether the resonant system is undermatched or overmatched to the external load.

621.317.616† : 681.85 **401**

The Variable-Disk-Speed Method of Measuring the Frequency Characteristics of Pick-Ups.—Terry. (*See* 275.)

621.317.7.001.4 **402**

Operation and Care of Circular-Scale Instruments: Part 3 — Electrodynamic Type Instruments.—J. Spencer. (*Instruments*, Sept. 1948, Vol. 21, No. 9, pp. 836–839; 852.) Discussion of single-phase and polyphase wattmeters and frequency meters, with special reference to the Westinghouse Type KF-24 and the General Electric Type AB-12. Parts 1 & 2: 146 of January.

621.317.7.029.64 : 621.392.26† **403**

A Michelson-Type Interferometer for Microwave Measurements.—B. A. Lengyel. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1242–1244.) 1949 I.R.E. National Convention paper noted in 1713 of 1949 (No. 14). The optical Michelson interferometer is modified by replacing one of its branches by a directional coupler and a waveguide. Various applications are discussed. An instrument for λ 3.2 cm is described. For a similar instrument, see 162 of January (Pippard).

621.317.74 : 621.396.9 **404**

A Radar Test Set for the Super-High-Frequency Band of 9 000–9 700 Mc/s.—W. Rosenberg, J. S. Fleming & E. D. Hart. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 476–482.) The quantities such a test set must be able to measure are: (*a*) mean power output of transmitter, (*b*) mean transmitter frequency, (*c*) s.w.r. in the aerial feeder, (*d*) width of the transmitter spectrum, (*e*) recovery time of the t.r. switch, (*f*) receiver sensitivity, (*g*) the i.f. response curve. Facilities must also be provided for checking receiver a.f.c. and for tuning. The waveguide system and circuits of a set meeting these requirements are described, and the methods of making the various measurements are outlined.

621.317.755 **405**

High-Frequency Oscillography.—R. Theile. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 241–264. In German.) The development of high-power sealed c.r. tubes is described and some details of the A.E.G. tube (noted in 1946 of 1943) are given. Hollmann's microwave oscillograph (1977 of 1940) and transit-time oscillography using dynamic Lissajous figures (544 of 1940) are discussed and also single-sweep methods (2198 of 1941). References to 33 relevant German papers are given.

621.317.755 : 621.317.761 **406**

Frequency Spectrometer.—W. Kroebel. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 227–233. In German.) A method is described for analysing the signals from a transmitter and displaying the frequency spectrum on a c.r.o. The method can be used throughout the range from the longest waves to cm waves.

621.317.755.087.4 **407**

A Miniature Portable Cathode-Ray Oscillograph Recorder.—C. F. Johnson. (*Instruments*, Sept. 1949, Vol. 22, No. 9, pp. 800–801.) A 1-in. c.r.o. Type RCA-913 is fixed inside a 3-in. brass tube which is silver-soldered to the top of a light-tight sheet-metal box of dimensions 2 × 2.5 × 5 in., containing the readily removable paper-drive mechanism. A tape mask forming a 1/64-in. slit is placed over the face of the c.r.o., and the intensity and size of the spot are adjusted to give the correct exposure. For a given spot intensity, the width of the trace is inversely proportional to the speed of the spot. A dry battery supplies power to the c.r.o. and to an associated high-gain a.f. amplifier.

621.317.761 **408**

Absolute Frequency Measurement.—A. Scheibe. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 216–226. In German.) An account of the quartz clocks, frequency and wavelength standards, oscillation generators and associated magnetron valves, and indicators for measurements with cm and mm waves, of the Physikalisch-Technische Reichsanstalt in the years 1936 to 1945.

621.317.78 **409**

Power Meter for Communication Frequencies.—R. L. Linton, Jr. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1245–1246.) A portable, rugged instrument which is easy to use, for measuring the power delivered to an aerial in the frequency range 2–20 Mc/s and power range 1–200 W. Accuracy is within ±50%. A directional coupler loop of the type described by Early (1007 of 1947) is used.

621.317.79 : 621.396.61 410

Signal Generators.—C. Schmelzer. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*. Part 2, 1948, pp. 211–216. In German.) Sources are considered which will provide oscillations approximating closely to sine waves over a wide range of both frequency and amplitude. References to 27 relevant German papers are given.

621.385.001.4 411

A Universal Visual Valve Tester.—F. L. Hill & C. W. Brown. (*Electronic Engng.*, Nov. 1949, Vol. 21, No. 261, pp. 425–430.) The valve tester described incorporates a c.r. tube on which is displayed a complete family of V_a I_a curves for ten different grid voltages. Sockets are provided to accommodate all modern types of valve, and a universal connection board enables any electrode to be connected to any supply. The 12-phase generator used is operated from 3-phase or single-phase mains.

621.392.26† : 621.396.662 412

Corrections to the Attenuation Constants of Piston Attenuators.—Brown. (*See* 280.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.768 : 549.514.51 413

A Piezoelectric Device for Measuring Momentary Accelerations.—E. Völcker. (*Frequenz*, Aug. 1949, Vol. 3, No. 8, pp. 244–249.) Details of the construction and calibration of apparatus in which the acceleration to be measured is applied to a small cylinder whose slight motion compresses a quartz rod. The resulting potential developed on the side of the rod is amplified and applied to one pair of deflection plates of a c.r.o.

534.321.9 414

Ultrasonics: A Brief Survey.—Jupe. (*See* 269.)

534.321.9.001.8 : 534.88 415

Obstacle Detection using Ultrasonic Waves in Air.—G. Bradfield. (*Electronic Engng.*, Dec. 1949, Vol. 21, No. 262, pp. 464–468.) Discussion of experimental results obtained with a spark transmitter using a standard 14-mm spark plug with the 1-mm gap at the focus of a 7½-in. paraboloid; the microphone included a hollow double bimorph Rochelle-salt crystal of dimensions 9.5 × 9.5 × 3 mm, resonant at about 19 kc/s, and was used with a Type-CV138 head amplifier. Frequencies of 52 and 110 kc/s were used. Signal strengths obtained from various objects at different distances are tabulated. The directivity of the apparatus was very high, giving good resolution of objects which are side by side; poor scattering surfaces such as walls gave feeble returns, especially at oblique incidence. At this stage of development, ultimate success with equipment of this type as an obstacle detector for use by the blind, or as a fog safety device, cannot be confidently predicted.

620.91 : 537.56 416

Comparative Survey of Ion Guns: Parts 1 - 3.—M. Hovaux & I. Dujardin. (*Nucleonics*, May–July 1949, Vol. 4, Nos. 5 & 6, pp. 7–9, 12–29 & Vol. 5, No. 1, pp. 67–71.) Part 1: General properties and specific characteristics of typical guns. Part 2: Properties and relative merits of representative ion sources, with bibliography of 86 references. Part 3: Necessary refinements of ion gun design.

621.317.39 : 531.77 417

Precision Speed Measurement of Rotating Equipment.—M. W. Hellar, Jr. (*Gen. elect. Rev.*, Oct. 1949, Vol. 52, No. 10, pp. 22–26.) An instrument is described which will measure instantaneous values of speed of rotation with accuracies of ± 0.025% and ± 0.125% over the ranges 1 710–1 780 r.p.m. and 1 350–1 700 r.p.m. respectively. A signal-frequency voltage is derived from the machine under test by a generator loading device or by a photocell method, and mixed with a standard-frequency voltage generated by a synchronous motor driving a small induction alternator. The difference frequency is selected and measured to give slip speed frequency is selected and measured to give slip speed in r.p.m., from which running speed is at once determined.

621.317.39 : 620.172.222 418

The Measurement of Changes in Length with the aid of Strain Gauges.—A. L. Biermasz & H. Hoekstra. (*Philips tech. Rev.*, July 1949, Vol. 11, No. 1, pp. 23–31.) Description of strain gauges Types GM4472 and GM4473 manufactured by Philips, and associated bridges. Various applications are considered.

621.365.54† 419

Radio-Frequency Induction Furnaces.—(*Metallurgia, Manchr*, Oct. 1949, Vol. 40, No. 240, p. 336.) Brief illustrated description of a Metropolitan-Vickers 25-kW 600-kc/s unit which can be used to melt 20 lb of ferrous metal in about 30 min, using 380–460-V 50-c/s 3-phase power supply. Similar 5-kW and 10-kW units are also mentioned.

621.365.54† 420

High-Frequency Induction Heating.—(*Metallurgia, Manchr*, Oct. 1949, Vol. 40, No. 240, pp. 332–334.) Description of a new forging shop at John Garrington & Sons, Bromsgrove. Three 10-kc/s generators are installed, each consisting of two 150-kW units driven by a common motor and feeding separate induction heaters for small billets. There are also three 3-kc/s generators, each with two 250-kW units which can be run independently or in parallel, for heating larger billets. Power consumption varies from 400 to 500 kWh per ton according to the size of billet. By using induction heating, high rates of production can be maintained in clean and airy buildings. Forging can be begun within a few minutes of starting up.

621.365.54† 421

Induction Heating.—(*Metal Ind., Lond.*, 9th Dec. 1949, Vol. 75, No. 24, pp. 498–501.) Illustrated description of various applications in the non-ferrous metal industry.

621.365.54† : 621.785.6 422

Multi-Purpose Induction Hardening Units.—Please alter title of 188 of January to read as above.

621.365.55† 423

Dielectric Heating: Applications in the Foundry.—J. Pound. (*Metal Ind., Lond.*, 21st Oct.–4th Nov. 1949, Vol. 75, Nos. 17–19, pp. 351–353, 379–381 & 399–400.) The use of dielectric heating for the baking of resin-bonded sand cores is considered.

621.38.001.8 424

Industrial Applications of Electronic Techniques.—H. A. Thomas. (*Proc. Instn elect. Engrs*, Part I, Nov. 1949, Vol. 96, No. 102, pp. 323–324.) Discussion on 3992 of 1947.

621.38.001.8 : 578.088.7 425

Biological Properties of Microwaves.—L. de Seguin. (*Onde élect.*, Oct. 1949, Vol. 29, No. 271, pp. 368–377; *Ann. Radioélect.* Oct. 1949, Vol. 4, No. 18, pp. 331–343.) Experiments were conducted to determine the effect of microwaves on bacteria, tissue growth and capillary circulation, and their penetration of dead and living tissue. Microwaves have much greater penetration than infra-red rays and are preferable to the longer e.m. waves for therapeutic heat treatment.

621.38.001.8 : 786.6 **426**

The Hammond Spinet.—A. Douglas. (*Electronic Engng.*, Dec. 1949, Vol. 21, No. 262, pp. 461–463.) General description and circuit details of a simplified type of instrument for home use, similar to the organ noted in 470 of 1949 (Wells) and back references.

621.383 : 551.576 **427**

Telemetry of Clouds by means of Light Pulses.—A. Baude. (*Radio franç.*, Oct. 1949, No. 10, pp. 3–17.) A very short pulse of light is sent vertically upwards from a source at the focus of a parabolic mirror; the light scattered from the base of a cloud towards a receiving photocell near the transmitter is concentrated by a second parabolic mirror. Electronic methods are used to measure the time interval t between the transmitted and received pulses; the cloud height is at once given by $h = 150t$, where t is in μs. Measurement accuracy to a fraction of 1 μs is necessary. Equipment details of two French sets, with typical records, are given. See also 1943 of 1946 (Moles) and 2657 of 1946.

621.384.6 : [615.849 + 53 **428**

The Development of Linear Accelerators and Synchrotrons for Radiotherapy and for Research in Physics.—J. Cockcroft. (*Proc. Instn elect. Engrs*, Part I, Nov. 1949, Vol. 96, No. 102, pp. 296–303.) A general survey, with brief reference to early types, the travelling-wave accelerator, the Atomic Energy Research Establishment accelerator, the 10-MeV accelerator developed by Metropolitan-Vickers for the Medical Research Council, and a 30-MeV betatron-started synchrotron. Applications to radiotherapy and physical research are considered. See also 175–177, 1148 and 2595 of 1949.

621.384.611.1† **429**

Electronics applied to the Betatron.—T. W. Dietze & T. M. Dickinson. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1171–1178.) Discussion of circuits for electron injection and ejection, and X-ray monitoring.

621.384.611.2† : 621.396.611.4 **430**

Quarter-Wavelength Coaxial-Line Resonators for Betatron-Started Synchrotrons.—F. K. Goward, J. J. Wilkins, L. S. Holmes & H. H. H. Watson. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 508–516.) Description of air-spaced and silvered-dielectric resonators for use in synchrotrons giving energies < 400 MeV. Methods of feeding, monitoring and tuning, modes of resonance, power requirements, and measurements on resonators which have been incorporated in 8-MeV and 30-MeV synchrotrons are considered.

621.385.833 **431**

The Design and Construction of a New Electron Microscope.—M. E. Haine. (*Proc. Instn elect. Engrs*, Part I, Nov. 1949, Vol. 96, No. 102, pp. 303–304.) Discussion on 2041 of 1948.

621.385.833 : 016 **432**

Metallurgical Achievements of the Electron Microscope.—G. A. Geach. (*Metallurgia, Manchr*, Oct. 1949, Vol. 40, No. 240, pp. 319–324. Bibliography, pp. 324–326.) A review of the literature to the end of 1948.

621.385.833 : 669.017 **433**

Electron Microscope and Diffraction Study of Metal Crystal Textures by means of Thin Sections.—R. D. Heidenreich. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 993–1010.)

621.365.5 **434**

Radio-Frequency Heating Equipment. [Book Review]—L. L. Langton. Publishers: Pitman & Sons, London, 196 pp., 17s. 6d. (*Metal Ind., Lond.*, 9th Dec. 1949, Vol. 75, No. 24, p. 496.) "... the time is suitable for such a book on the theory and practice of the generator. This book covers the subject in detail and is one that should be on the list of all engineers interested in this field of electronics."

621.38.001.8 **435**

Electronics in the Factory. [Book Review]—H. F. Trewman (Ed.). Publishers: Pitman & Sons, London, 183 pp., 20s. (*Electronic Engng.*, Dec. 1949, Vol. 21, No. 262, pp. 475–476.) "Compiled by members of the staff of Electrical and Musical Industries, Ltd ... but the contents are well representative of all branches and manufacturers .... The writing is clear and not too tedious to be read by the directors, general managers and production engineers to whom the book is addressed." Subjects covered include timing circuits, counting, motor control, regulation, heating, servomechanisms, photocell applications and strain gauges.

PROPAGATION OF WAVES

538.566 **436**

On the Boundary Conditions in the case of Two Absorbent Media in Contact.—H. Arzeliès. (*Ann. Phys., Paris*, Nov./Dec. 1948, Vol. 3, pp. 637–654.) The general boundary conditions for such media are derived from Maxwell's equations. The results can be used to express the theory of reflection in a very simple form. See also 3250 of 1947 (Booker).

538.566.2 : 621.396.67 **437**

The Magnetic Dipole in a Stratified Atmosphere.—Eckart. (*See* 284.)

621.396.11 **438**

Theory of Wave Propagation over the Earth, including the Influence of the Troposphere.—W. Pfister. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 127–133. In German.) The results of investigations on wave propagation carried out before the war are mostly to be found in three books: (*a*) Lehrbuch der drahtlosen Nachrichtentechnik, Band 2; Ausstrahlung, Ausbreitung und Aufnahme elektromagnetischer Wellen, by H. Lassen (Springer, 1940). (*b*) Die Ausbreitung der elektromagnetischen Wellen, by B. Beckmann (Akad. Verl. Ges., 1940). (*c*) Vol. 1 of Fortschritte der Hochfrequenztechnik (Akad. Verl. Ges., 1941), which includes contributions by H. Lassen, J. Grosskopf and B. Beckmann. References to 15 later papers are given.

621.396.11 : 551.510.535 **439**

Theory of Wave Propagation in the Ionosphere.—W. Becker. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 134–142. In German.) Discussion of investigations on (*a*) partial reflection by ionosphere layers, (*b*) the validity of the ray theory, (*c*) the perpendicular incidence of plane e.m. waves on a plane layered ionized medium under the influence of a magnetic field, (*d*) the propagation of e.m. waves in a layered ionized medium under the influence of a magnetic field, for oblique incidence, (*e*) indirect signals and transmission of s.w. signals round the earth. References to 13 relevant papers are given.

621.396.11 : 551.510.535 **440**

Thermal Expansion of Ionospheric Layer and Temporary Morning Disappearance of Radio Signals.—S. S. Banerjee & R. N. Singh. (*Nature, Lond.*, 26th Nov. 1949, Vol. 164, No. 4178, p. 925.) The reception in India of signals via the ionosphere, at wavelengths of the order of 20 m, often ceases for an hour or more after sunrise.

It is suggested that this is due to a thermal expansion of the F_2 layer which is more than sufficient to counterbalance the increase in ionization as the sun's altitude increases.

441

621.396.11 : 621.317.3

Wave-Propagation Measurements.—B. Beckmann. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 143–167. In German.) Review of investigations concerning ground waves, propagation mechanism, angle of incidence, polarization, scattering, attenuation, and of u.s.w. measurements of tropospheric propagation, with references to 54 relevant German papers.

442

621.396.812.3 : 551.510.535

Statistical Analysis of Fading of a Single Downcoming Wave from the Ionosphere.—S. N. Mitra. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 505–507.) The fading of a single magneto-ionic component of a radio wave of frequency 2–6 Mc/s, incident vertically on the ionosphere and reflected downwards, was recorded at two points 100 m apart. The records which indicate that the fading was not due to a regular ionospheric drift are analysed. The results agree, to the first order, with the assumption that in the reflecting region there are irregularities moving with velocities in the line of sight having a Gaussian distribution with r.m.s. value about 2–3 m/sec. Possible ionospheric causes for a significant discrepancy between the records and the simple theory are discussed. There is some experimental evidence to suggest that the irregularities responsible for the fading are situated below the E-region reflection point for 4-Mc/s waves. See also 96 of January and 443 below.

443

621.396.812.3 : 551.510.535

The Fading of Downcoming Radio Waves of Medium and High Frequencies.—R. W. E. McNicol. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 517–524.) The production of fading is considered in terms of an ionosphere with irregularities varying horizontally. The theoretical nature of the fading curve, with regard to both the distribution of amplitude and the variation with time, is indicated and compared with experimental results. The relative magnitudes of the steady and random components of a wave exhibiting fading were determined for different conditions; the results are summarized. From the rate of change of amplitude it is possible to calculate the effective velocity of the irregularities in the ionosphere, assuming that the fading is caused either by turbulent motion of the irregularities, or by a steady drift of the irregular ionosphere as a whole. The effective velocities deduced from 122 records made for vertical incidence and 55 for oblique incidence, lie between 0.3 and 8.0 m/sec, with an average value of 1.9 m/sec. See also 442 above.

444

621.396.11 : 621.396.813

An Analysis of Distortion Resulting from Two-Path Propagation.—Gerks. (*See* 451.)

RECEPTION

445

621.396.621

Ekco Model CR61.—(*Wireless World*, Dec. 1949, Vol. 55, No. 12, pp. 480–482.) Test report on an all-wave car receiver with permeability tuning.

446

621.396.621 : 621.396.619.11 : 13

A Simple Frequency Discriminator for A.M.-F.M. Receivers.—E. G. B. (*Philips tech. Commun., Aust.*, 1949, Nos. 2/3, pp. 17–22, 27.)

447

621.396.621 : 621.396.619.11 : 13 : 621.396.615

Signal Frequency and Oscillator Circuits for A.M.-F.M. Receivers.—E. G. Beard. (*Philips tech. Commun., Aust.*, 1949, Nos. 2/3, pp. 23–27.) The input and oscillator circuits of a medium-wave receiver can be modified at small cost for f.m. reception. Small inductors inserted in the signal-grid and oscillator-grid leads are 'series-tuned' by the normal medium-wave 350-pF tuning capacitors. A cathode choke is inserted and switching is effected by short-circuiting the medium-frequency coils. Tests show that the modification is effective while only one or two stations are operating on the m.f. band. Details of modifications of the remaining a.m. receiver circuits will be given later.

448

621.396.621 : 621.396.619.13

The Demodulation of a Frequency-Modulated Carrier and Random Noise by a Discriminator.—N. M. Blachman. (*J. appl. Phys.*, Oct. 1949, Vol. 20, No. 10, pp. 976–983.) The discriminator is regarded as consisting of two selective circuits, both fed by the output of the i.f. amplifier but peaked at different frequencies, feeding rectifiers whose outputs are subtracted. The effect of passing random noise through each of these circuits is considered by Rice's method (440, 2168 and 2169 of 1945) with due regard to the correlation between the noise voltages fed to the two rectifiers. Quadratic and linear rectification are considered. The results are applied to the case of a rectangular i.f. noise spectrum, and the signal/noise ratio is determined for the cases of narrow-band and wide-band f.m. The results are very much like those for Middleton's idealized representation of the discriminator (2619 and 3532 of 1949), and are tabulated with corresponding a.m. results. The optimum signal/noise ratio for narrow-band f.m. without a limiter occurs when the discriminator is designed for the least possible bandwidth; this optimum ratio differs very little from that for a.m.

449

621.396.621 : 621.396.645.371

Some Dangers in the Use of Negative Feedback in Radio Receivers.—E. G. Beard. (*Philips tech. Commun., Aust.*, 1949, Nos. 2/3, pp. 3–13.) A non-mathematical discussion of the principles of negative feedback, with criticism of some popular feedback circuits in audio amplifiers. Wrong positioning of the feedback resistor or volume control in the circuit can increase hum or lead to anomalies in amplifier gain. Recommended general-purpose feedback circuits are shown. Positive envelope feedback can occur by accident in a negative-feedback circuit; a circuit is described in which this is avoided. Distortion in a reflex receiver circuit due to an unwanted envelope feedback effect, and methods of circumventing this, are discussed from a practical point of view.

450

621.396.622.7 : 621.385.5 : 621.396.619.13

The 'ϕ Detector', A Detector Valve for Frequency Modulation.—Jonker & van Overbeek. (*See* 505.)

451

621.396.813 : 621.396.11

An Analysis of Distortion Resulting from Two-Path Propagation.—J. H. Gerks. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1272–1277.) For a.m., nonlinear distortion caused by two-path propagation is a result of over modulation in the resultant signal. This distortion becomes severe only when the time delay on the secondary path is large and the amplitudes are nearly equal. For f.m., the instantaneous frequency of the resultant signal has spike-shaped variations which reach large amplitude when the signals are nearly equal. When the discriminator is designed to respond linearly to a very wide frequency deviation, an averaging process takes place in the receiver which tends to minimize distortion; a discriminator range of several megacycles per second may be necessary for optimum reduction of distortion.

452

621.396.821 : 551.594.9

Atmospherics.—Siedentopf. (*See* 361.)

621.396.828 453

Suppressing Impulse Noise.—D. C. Rogers. (*Wireless World*, Dec. 1949, Vol. 55, No. 12, pp. 489–492.) The duration of impulse noise is nearly always substantially less than 1 μs, and the impulses seldom overlap. A circuit is described in which this short duration is used to distinguish noise from signal, whatever their relative amplitudes. The output from the receiver detector is divided into two parts, one passing through a phase inverter and an attenuator, the other through a high-pass filter and a pulse shaper ; these parts are then re-combined to form a noise-free resultant. The principal use of the circuit is for frequencies above 30 Mc/s and it is restricted to cases where selectivity is unimportant; it can thus be used for suppressing ignition interference in the sound section of a television receiver, but not with broadcast or communication receivers.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 : 621.396.4 : 621.396.619 454

Band Width and Transmission Performance.—C. B. Feldman & W. R. Bennett. (*Bell Syst. tech. J.*, July 1949, Vol. 28, No. 3, pp. 490–595.) A general and very comprehensive discussion is given of the relation between transmission bandwidth, power, noise, interference and overall performance for various multiplex systems, with due attention to all the relevant practical factors. Frequency division and time division are considered in combination with either a.m. or f.m. of a carrier by pulse-position, pulse-amplitude or pulse-code signals. The advantages and disadvantages of trading bandwidth for improved transmission are discussed in detail. The superiority of p.c.m. arising from the facility of signal regeneration, coding and power reduction is stressed, and the advantages of p.c.m./a.m. over f.m. for long television relay routes are noted.

621.395 : 061.3 455

C.C.I.F. Meetings, May and July, 1949.—(*P. O. elect. Engrs' J.*, Oct. 1949, Vol. 42, Part 3, pp. 168–171.) Brief summaries are given of work done by committees on protection, corrosion, long-distance transmission, local transmission, signalling, switching and symbols.

621.395.44 : 621.395.645.37 456

Transmitting Amplifier for the K2 Carrier System.—Fleming. (*See* 314.)

621.395.47 457

Analysis-Synthesis Telephony, with special reference to the Vocoder.—R. J. Halsey & J. Swaffield. (*Proc. Instn elect. Engrs*, Part III, Nov. 1949, Vol. 96, No. 44, pp. 497–504.) Discussion on 513 of 1949.

621.396.1 : 061.3 458

The High Frequency Broadcasting Conference, Mexico City, October 1948-April 1949.—(*P.O. elect. Engrs' J.*, Oct. 1949, Vol. 42, Part 3, pp. 166–168.) The detailed allocation of frequencies within the bands scheduled at the Atlantic City conference was discussed.

621.396.619.16 459

Pulse Modulation.—E. M. Deloraine. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 222–227.) Reprint. See 2636 of 1949.

621.396.619.16 460

Signal-to-Noise Ratio Improvement in a Pulse-Count-Modulation System.—A. G. Clavier, P. F. Panter & W. Dite. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 257–262.) Reprint. See 2325 of 1949.

621.396.619.16 461

Theoretical Study of Pulse-Frequency Modulation.—A. E. Ross. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1277–1286.) The accuracy in reconstruction of the signal depends not only on the number of sampling points per period but also on their distribution. Assuming a fixed average pulse frequency (sampling rate), the maximum range of the signal frequency for which the corresponding periodic signal may be transmitted with acceptable accuracy is determined. P.f.m. is as far as possible considered independently of the circuit used to realize it. The sampling for certain ranges of the ratio of signal frequency to pulse frequency is considered specially. The behaviour of the general periodic signal cannot be deduced from that of its sinusoidal components; the methods here developed must be applied directly to the particular waveform considered. An example is given.

621.396.619.16 : 621.396.41 462

A Technique for the Design of Pulse Time Multichannel Radio Systems.—M. M. Levy. (*J. Brit. Instn Radio Engrs*, Nov. 1949, Vol. 9, No. 11, pp. 380–411.) The transmitter uses a delay line in a special feedback circuit, and a square wave. Each channel is selected by a correct tapping on the delay line, and the channel boundaries are accurately timed by the square wave. The whole system is automatic. Special 3-valve 'power multivibrators' are used as generators which produce pulses of high peak power in impedors of low value; a shock-excited tuned circuit is used to define the length of the pulse. Trapezoidal pulses trigger these generators. When a signal pulse is added, the instant of triggering is modulated in time. The use of trapezoidal pulses ensures that the pulse time modulation never crosses the boundaries of the channel.

The demodulation process consists of transforming the time modulation into pulse length modulation by means of another type of multivibrator which is sensitive only during the time allocated to the corresponding channel, and which triggers only when the corresponding channel pulse appears. The demodulated signal is filtered with a low pass filter. A single circuit is used for several channels; this reduces the number of valves required to about 3 per 2 channels, including mixing valves. The mixer circuit comprises cathode followers with cathodes connected together. Crosstalk is negligible if 'power multivibrators' are used. The optimum number of channels seems to be 20–40, and the maximum number 100. Block and simplified circuit diagrams are given for a 24-channel system, and an experimental 20-channel system built in 1943 is described which was satisfactorily tested for distances up to about 36 miles in mobile vans. See also 3054 of 1947 and 2620 of 1948.

621.396.65 : 621.397.5 463

Television Radio Relay: London-Birmingham Link.—(*See* 471.)

621.396.931 464

V.H.F. Radio Equipment for Railways and Heavy Industries.—(*Engineer, Lond.*, 28th Oct. 1949, Vol. 188, No. 4892, p. 502.) A brief description of a specially designed 15–20-W f.m. transmitter and receiver. The whole is mounted in a light-alloy dustproof and weatherproof case and can be installed in any position. A built-in selective-calling device using uniselectors makes constant monitoring of the mobile units unnecessary, and is applicable to simplex, two-frequency simplex or duplex working. A 3-digit system will cover 500 substations. The control unit is in the form of a telephone handset. Provision is made for mains or battery operation and for the substitution of a.m. units if desired.

SUBSIDIARY APPARATUS

621.314.63 465

Image Force in Rectifiers.—P. T. Landsberg. (*Nature, Lond.*, 3rd Dec. 1949, Vol. 164, No. 4179, pp. 967–968.) A theoretical characteristic which allows for the effect

of image force was given by Mott (4136 of 1939). A general method of allowing for this force is given, and applied to the Cu_2O rectifier.

621.315.59 : 537.311.4 **466**

On the Theory of the A.C. Impedance of a Contact Rectifier.—Bardeen. (*See* 349.)

621.316.721 **467**

A Stabilized 100-A Power Supply.—D. E. Caro & J. K. Parry. (*J. sci. Instrum.*, Nov. 1949, Vol. 26, No. 11, pp. 374–377.) "A system is described for maintaining constant currents of up to 100 A through highly inductive loads. The system stabilizes against changes in mains voltage and generator characteristics, reducing the short-period current variations to less than 1 part in 2 000. Current changes due to load-impedance variations are reduced by a factor of approximately 20. The main interest in the system is the electromagnetic method of comparing the load current with a standard current."

621.396.682 : 621.316.722 **468**

Heater Compensation improves Stabilized Power Supplies.—(*Tech. Bull. nat. Bur. Stand.*, Oct. 1949, Vol. 33, No. 10, pp. 115–116.) A method developed by R. C. Ellenwood & H. E. Sorrows. Heater-voltage fluctuations are used to compensate for line-voltage fluctuations. The method can be applied to power supplies using degenerative voltage stabilizers in which the output voltage is compared with a fixed reference voltage, the difference voltage being used to alter the resistance of a control valve. Circuit and component details are given. For another account see *Radio & Televis. News, Radio-Electronic Engng Supplement*, Nov. 1949, Vol. 13, No. 5, pp. 8, 30.

621.396.682 : 621.319.3 **469**

Stabilized High Voltage Power Supply for Electrostatic Analyzer.—R. L. Henkel & B. Petree. (*Rev. sci. Instrum.*, Oct. 1949, Vol. 20, No. 10, pp. 729–732.) A description of an a.c. operated power supply which provides a voltage continuously variable from 5 to 50 kV with stability and accuracy of measurement within ± 0.01%. The stabilizer is of the usual degenerative type, but the error signal is amplified by a photocell galvanometer, d.c. amplifier and applied to a series impedance in the h.v. transformer primary circuit. The generator beam energy automatically follows adjustments of a l.v. potentiometer used for voltage measurements in this supply.

771.3 : 621.3.087.5 : 551.510.535 **470**

Continuous [-recording] Oscillograph Camera for Ionosphere Measurements.—J. E. Hacke, Jr. (*Instruments*, Oct. 1948, Vol. 21, No. 10, pp. 914–915.) Illustrated description, with special reference to the mechanism for moving the photographic paper.

TELEVISION AND PHOTOTELEGRAPHY

621.396.65 : 621.397.5 **471**

Television Radio Relay: London-Birmingham Link.—(*Wireless World*, Dec. 1949, Vol. 55, No. 12, pp. 474–476.) The link has four unattended relay stations. It provides a single vision channel which can be used in either direction. The installation is permanent and designed for reliability; all apparatus is duplicated. Transmission frequencies of 870 Mc/s and 890 Mc/s are used. At the terminal station the video-frequency signal is used for f.m. of a 34-Mc/s carrier; this modulated carrier modulates a 904-Mc/s carrier and the sideband in the range 868.5–871.5 Mc/s is fed to the aerial. Relay-station transmitter output is about 10 W and the gain is about 70 db. Dipole aerials with reflectors are used; their gain is about 28 db. A special filter provides 70-db discrimination between wanted and unwanted frequencies in the a.m. process; it uses coaxial resonant circuits in which the conductors are separated by alternate short sections of air and polythene, so that the total length of the filter is reduced from 10 ft to 18 in. A r.f. switch is used for the coaxial circuits of the transmitters and receivers and for bringing in duplicate equipment when a fault occurs. It has no contacts and operates by moving plungers into coaxial stubs, so that in either extreme position an infinite impedance is presented on one side and a zero impedance on the other. The quality of the picture originating in London and seen in Birmingham is as good as that usually obtained near London. See also 297 of 1948, 1279 and 1857 of 1949 (Stanesby & Weston), and *Electronic Engng*, Dec. 1949, Vol. 21, No. 262, pp. 457–460.

621.397.5 **472**

Television (including Scanning).—F. Schröter. (*FIAT Review of German Science, 1939–1946; Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 185–210. In German.) The principal developments in fundamental investigations in Germany during the war years were concerned with improvement of c.r. tube spot sharpness, reduction of errors in the deflection system, sensitive photoelectric layers and secondary-emission layers with a high multiplication factor, with applications to picture-scanning tubes, projection apparatus and colour television. These questions are reviewed and also improvements in technique and in transmitting and receiving equipment. References to 53 relevant German publications are included.

621.397.5 : 621.315.212 **473**

London-Birmingham Television Cable.—H. Stanesby & W. K. Weston. (*Elect. Commun.*, Sept. 1949, Vol. 26, No. 3, pp. 186–200.) See 1279 and 1857 of 1949.

TRANSMISSION

621.396.61 **474**

Power Tubes in Parallel at U.H.F.—J. R. Day. (*Electronics*, Nov. 1949, Vol. 22, No. 11, pp. 166–170.) An arrangement in which the power output is proportional to the number of valves in parallel and the upper frequency limit is the same as for each individual valve used alone. The active input and output circuits comprise 3 coaxial cylinders and 4 shorting rings. The valves, which must be either of the planar type or have an external disk connection to the separation electrode, are arranged with their axes lying in a plane normal to the cylinders and meeting the cylinder axis. The plane of the valves is about half way between the shorting rings, so that they are, in effect, at or near the voltage loop of a resonant half-wave coaxial circuit. Spurious modes presented no difficulty. An alternative rectangular arrangement with valves on opposite sides is briefly considered.

621.396.61 **475**

Miniature Tubes in a Band-Switching Exciter.—W. Mayer. (*QST*, Dec. 1949, Vol. 33, No. 12, pp. 11–15.) Description of a 75-W variable-frequency driver unit, built for eliminating television interference, contained in one compact shielded unit, with complete band-switching for amateur frequencies.

621.396.61 **476**

TVI on 160 Meters?—P. S. Rand. (*CQ*, Dec. 1949, Vol. 5, No. 12, pp. 11–14, 62.) Circuit diagram and component details for an amateur transmitter free from television interference. Part of the circuit was also used for the 10-m transmitter discussed in 1530 of 1949.

621.396.61 : 621.396.712 **477**

KTBS' New Transmitter.—W. M. Witty. (*Broadcast News*, Sept. 1949, No. 56, pp. 46–53.) A general illustrated description. A 6-element directional aerial array

is used. Special precautions against possible flooding, lightning etc. are discussed. The power is 10 kW during the day and 5 kW at night, with a frequency of 710 kc/s.

621.396.619.22 478

Non-Linear Inductance and Capacitance as Modulators for Amplitude-Modulation Systems.—D. G. Tucker. (*P.O. elect. Engrs' J.*, Oct. 1949, Vol. 42, Part 3, pp. 156–159.) The performance of modulators using the nonlinear characteristics of suitable inductors and capacitors is discussed theoretically. The output of any sideband depends upon its frequency, and low-frequency products have a small amplitude. Practical circuits and measurement results are given for a magnetic modulator using a nonlinear inductor.

621.396.619.23 479

New Modulator Circuit utilizes 807's in Class B with Zero Bias.—A. M. Seybold. (*Radiotronics*, July/Aug. 1949, No. 138, pp. 64–65.) A 20 000-Ω 1-W resistor connects the control grid to the screen grid of each of the Type-807 valves, the cathodes are earthed and the secondary of the driver output transformer is connected to the two screen grids, the centre tap being also earthed. With a supply voltage of 750 V and peak grid-to-grid input of 555 V, audio output into a 6600-Ω load is 120 W. Type-2A3 valves are recommended for the driver. A circuit diagram, with component details, and functional curves for the Type-807 valves for various signal voltages, are given.

621.396.619.231 480

Modulators for High Power Transmitters.—H. A. Teunissen. (*Commun. News*, June 1949, Vol. 10, No. 2, pp. 41–51.) Anode modulation with a class-B modulator is generally used for medium- and short-wave transmitters; it is more efficient than other possible systems mentioned. Efficiency has recently been improved by using feedback and a cathode-follower driver stage. Damping resistors are not then needed in the grid circuit, and no transformer is necessary in the chain of amplifier stages over which feedback is acting; it is difficult to build a.f. transformers without introducing a considerable phase shift which might endanger stability. The feedback factor should be as large as possible; the best solution is to feed from the primary of the modulation transformer back to the cathode of the first stage, so that a very low output impedance is obtained. A coupling arrangement in which the anode of the modulated amplifier is shunt-fed is advantageous. The application of these principles to the design of the modulator of a 100-kW transmitter is discussed, and experimental results obtained during the testing of the modulator of a 40-kW transmitter are considered.

VALVES AND THERMIONICS

537.291 + 538.691 481

On the Theory of Axially Symmetric Electron Beams in an Axial Magnetic Field.—A. L. Samuel. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1252–1258.) A type of electron beam is proposed in which the space-charge repulsive forces are balanced by magnetic focusing forces, so that the beam may be made as long as required without any change in its cross-section. The equations governing the existence of such beams are derived.

621.385 482

A Survey of Modern Radio Valves: Part 1 — Introduction.—H. Stanesby. (*P.O. elect. Engrs' J.*, Oct. 1949, Vol. 42, Part 3, pp. 117–118.) Introduction to a series of articles on basic principles common to most valves, and on valves for use in various frequency ranges up to 30 kMc/s. See also 483 below.

621.385 483

A Survey of Modern Radio Valves: Part 2 — The Physical Principles of Thermionic Valve Operation.—K. D. Bomford. (*P.O. elect. Engrs' J.*, Oct. 1949, Vol. 42, Part 3, pp. 118–123.) Phenomena common to most valves are surveyed. Primary and secondary emission, space charge, transit time, and the functions of various electrodes in conventional valves are discussed. See also 482 above.

621.385 484

Low-Distortion Power Valves.—G. Diemer & J. L. H. Jonker. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, pp. 385–390.) A survey is given of various low-distortion valve constructions. Two new constructions for pentodes are described which involve (*a*) alignment of the first and second grids, (*b*) introduction of additional focusing rods into the electrode system. By either means the second harmonic of a single-stage class-A pentode amplifier can be considerably reduced. The total distortion can be halved for an output up to about 25% of the static anode dissipation. The new valves have I_a/V_g characteristics that are practically linear in the neighbourhood of the normal operating point. See also 4110 of 1937 (Kleen).

621.385 : 538.691 485

Study of the Magnetic Focusing of Cylindrical [electron] Beams.—G. Convert. (*Bull. Soc. franç. Elect.*, Oct. 1949, Vol. 9, No. 97, pp. 550–558.) General equations are developed for the electron motion in a system with axial symmetry, consisting of an electron gun sending a beam along a drift tube maintained at a constant potential, with a uniform axial magnetic field. Cylindrical beams are considered for the two cases in which the rotational moment has or has not the same value for all the electrons. Undulations of the beam within the drift tube are discussed and the importance is stressed of proper adjustment of the magnetic field at the entrance of the tube by fitting suitable screens.

621.385.001.4 : 533.59 486

An Improved Method of Testing for Residual Gas in Electron Tubes and Vacuum Systems.—E. W. Herold. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 430–439.) An ionization-gauge method in which the ion current due to the residual gas is converted into a.c. by modulation of the ionizing electron stream, while stray currents are left relatively unmodulated. The sensitivity limit of 10^{-11} A, or 10^{-9} A without neutralization of undesired quadrature components, compares favourably with the sensitivity limit of 10^{-7} A for conventional d.c. ionization gauges. Some applications of the method are shown.

621.385.012 : 517.54 487

Potential Functions for a Thermionic Vacuum Tube.—E. C. Okress. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 850–856.) Potential functions are derived, by application of conformal transformations, for a system comprising two parallel metal plates at the same potential, with three flat parallel-wire grids evenly spaced between them. The wires in the central grid are orthogonal to those in the other two grids. The results are applied to the Western Electric Type 104D valve, the construction of which approximates to the theoretical system considered.

621.385.032.2 : 777.1 488

Electrodes for Vacuum Tubes by Photogravure.—M. P. Wilder. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, pp. 1182–1184.) An improved method which enables complicated designs to be cut with little effort; its chief utility is in the manufacture of precision parts for experimental valves.

489

621.385.032.212

Circuits for Cold Cathode Glow Tubes.—W. A. Depp & W. H. T. Holden. (*Elect. Mfg. N.Y.*, July 1949, Vol. 44, No. 1, pp. 92–97.) Discussion of fundamental operating characteristics, and typical circuits using these tubes for relays, impulse generators, pulse counting, and interlocking.

490

621.385.032.216

Some Properties of the Ba_2SiO_4 Oxide-Cathode Interface.—A. Eisenstein. (*J. appl. Phys.*, Aug. 1949, Vol. 20, No. 8, pp. 776–790.) The thickness of the interface, its effective specific electrical conductivity, and the interface voltage developed by the flow of emission current are considered. Methods of measuring these quantities and results obtained are discussed.

491

621.385.032.216

Electron Emission and Conduction Mechanism of Oxide-Coated Cathodes.—R. Loosjes & H. J. Vink. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, p. 884.) Experiments with oxide cathodes, showing that the curve of $\log a$ against $1/T$ has a bend at 750°K, suggest that there are two conduction mechanisms operating in parallel in the coating, one with a low activation energy which predominates below 750°K and one with a high activation energy which predominates above that temperature. This theory gives a simple explanation of Mahlman's results (2094 of 1949). A fuller account of these investigations will appear in *Philips Res. Rep.*

492

621.385.032.216

Microanalysis of Gas in Cathode Coating Assemblies.—H. Jacobs & B. Wolk. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1247–1251.) Gases evolved during degassing and activation of oxide cathodes are considered.

493

621.385.032.216 : 536.7

Application of Thermodynamics to Chemical Problems involving the Oxide Cathode.—A. H. White. (*J. appl. Phys.*, Sept. 1949, Vol. 20, No. 9, pp. 856–860.)

494

621.385.15

A Microwave Secondary Electron Multiplier.—M. H. Greenblatt. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, pp. 646–650.) The multipactor, a dynamic electron multiplier, consists of two parallel secondary-emission plates with an a.c. voltage across them which is adjusted so that the transit time across the gap, for electrons starting with zero velocity from one of the plates when the field is passing through zero, is half the period of the a.c. voltage. Primary and secondary electrons are thus made to oscillate between the plates and multiplication takes place. Phase relations between the electrons and the field are discussed. When the multipactor was used as a γ-ray detector and the a.c. voltage was obtained from 10-cm power, the rise time of the pulse obtained was calculated as 5×10^{-10} sec and measured to be $< 10^{-7}$ sec. The dead-time was about 5 μs.

621.385.18.032.213 : 621.396.615.12

495

The Generation of High-Frequency Oscillations by Hot-Cathode Discharge Tubes containing Gas at Low Pressure.—E. B. Armstrong & K. G. Emeleus. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 390–394.) Discussion of experimental results concerning the generation of plasma electron oscillations by the discharge from a hot filament through ionized Hg vapour, Ar and other gases. Oscillations are produced both with bare metal cathodes and with oxide cathodes. The vacuum wavelengths of the e.m. waves produced range from 5 cm to 2 m. Under favourable conditions 1% of the anode power can be converted into energy of oscillation in a coupled external circuit. Restricted regions of oscillating plasma appear to exist primary electron beams traversing these regions undergo v.m.

496

621.385.2

The High-Frequency Response of Cylindrical Diodes.—E. H. Gamble. (*Proc. Inst. Radio Engrs, W. & E.*, Oct. 1949, Vol. 37, No. 10, p. 1206.) Summary only. The investigation includes the case of large signals which have components varying with time through an amplitude comparable with that of the constant polarizing voltage. The problem is regarded as quasi-stationary; the solution is based on Poisson's equation, Newton's equation and the continuity equation. The method of integral equations is used. A method of successive approximations is devised for each specific form of applied voltage. Though the solutions are approximate they throw light on the inner mechanism of the cylindrical diode. The simplifying assumptions applicable to the cases of space-charge-limited and temperature-limited operation are mentioned. A first approximation to numerical solutions was obtained by means of an analogue computer; the solutions are shown graphically and satisfy the integral equations to an accuracy within 2%. For large signals, considerable modulation of the electric field and of the velocity and trajectory of the electrons occurs. In some cases, electrons are returned to the cathode or oscillate radially in the interelectrode space. For large electrode radii, the solutions reduce to those applicable to the planar diode.

497

621.385.2

Extension of the Planar Diode Transit-Time Solution.—N. A. Begovich. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1340–1344.) Llewellyn's small-signal theory for the plane parallel diode (552 of 1936 and 1669 of 1941) has been extended to include a closed-form second-order and third-order solution for complete space-charge operation. This solution includes terms not given by Benham's conservation-of-charge method (148 of 1939).

498

621.385.2 : 621.3.011

The Unit of Perveance.—G. D. O'Neill. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, p. 1295.) Perveance is defined as the constant G in the current/voltage relation $I = GV^{\frac{3}{2}}$ applicable to a space-charge-limited diode. It is suggested that the unit for G should be associated with the name of Langmuir.

499

621.385.2 : 621.396.822

Transit Time Correction Factor for Cylindrical Noise Diodes.—H. Ashcroft & C. Hurst. (*Proc. phys. Soc.*, 1st Oct. 1949, Vol. 62, No. 358B, pp. 639–646.) The noise generated by a diode is always less than it would be if the transit of the electrons were instantaneous. The factor by which it is reduced is tabulated for transit angles θ up to $12\frac{1}{2}$ radians and for ratios r of anode to cathode radius up to 20. The reduction is zero when $\theta = 7.498$ radians and $\log_e r = 1.246$.

500

621.385.2 : 621.396.822

Temperature-Limited Noise Diode Design.—R. W. Slinkman. (*Sylvania Technologist*, Oct. 1949, Vol. 2, No. 4, pp. 6–8.) A general discussion of design problems and of the more important characteristics of such diodes. Design and construction details are given for a diode with approximately 8 W total dissipation and maximum operating frequency 500 Mc/s. A 50-W tube is in the experimental stage.

501

621.385.3

Pencil-Type U.H.F. Triodes.—G. M. Rose, D. W. Power & W. A. Harris. (*RCA Rev.*, Sept. 1949, Vol. 10, No. 3, pp. 321–338.) 1949 I.R.E. National Convention paper noted in 1824 of 1949 [No. 141]. A new type of triode

is described which satisfies basic requirements of minimum transit time, lead inductance and internal capacitance. It is small, has good thermal stability and low heater wattage, and is suitable for mass production. A double-ended construction is used, with the rod-type anode and cathode connections extending outwards from the two sides of a control-grid disk. The internal elements are cylindrical and coaxial. Characteristics and performance data are included for 4 different valves of this type, designed for use as amplifiers and low-power oscillators. Typical applications are described.

621.385.38 502

Hydrogen Thyratrons.—J. Grolleau. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 522–524.) The advantages and disadvantages of hydrogen thyratrons are enumerated and a thyratron capable of handling high peak power is described. This valve, developed by the Compagnie Générale de Télégraphie sans Fil (C.S.F.), has an indirectly heated cathode in the form of a cylinder covered with fine-meshed Ni gauze to avoid flaking of the coating which usually occurs with large metal surfaces. Two thermal screens are arranged on either side of the cathode in order to maintain the cathode at its correct operating temperature of 820°C with the minimum heating current. The grid completely surrounds the anode/cathode space and the part which controls the damping of the arc consists of three baffles. Deionization is aided by keeping the anode/grid space reasonably small. The heater current is 11.5 A at 6.3 V, maximum peak voltage 16 kV, maximum peak current 160 A and mean current 100 mA. Pulses of duration up to 6 μs, with a recurrence frequency of 4 000 per sec, can be handled. Valve life is of the order of 500 hours. For peak powers higher than 1–2 MW, the thyratrons can be connected either in series or in parallel.

621.385.38 503

Extending Range D.C. Bias Control of Thyratron Plate Current.—L. Reiffel. (*Rev. sci. Instrum.*, Sept. 1949, Vol. 20, No. 9, pp. 699–702.) Advantages and disadvantages of three existing methods of controlling the average current in a thyratron with alternating anode voltage are discussed. In a new method, an alternating voltage is applied to the anode of the thyratron and another alternating voltage to the screen grid. The phase of the screen-grid voltage is shifted about 90° behind that of the anode voltage by a fixed phase-shifting network. An asymmetry is thus introduced into the critical grid bias control curve of the thyratron, and the minimum of this curve is shifted towards the 180° point of the anode voltage. Modifications of this simple system are discussed.

621.385.38 504

Hot-Cathode Thyratrons: Practical Studies of Characteristics.—H. de B. Knight. (*Proc. Instn elect. Engrs*, Part III, Sept. 1949, Vol. 96, No. 43, pp. 361–378. Discussion, pp. 379–381.) A survey of the main factors affecting the characteristics and life of thyratrons, based on experimental results. Such factors include electron emission from the cathode, ionization and current build-up, grid control, current-carrying capacity, the decay of ionization at the end of conduction, and the provision of suitable operating conditions. Other factors, such as positive-ion bombardment, may affect valve life but not electrical performance. The control characteristics of different types of grid are discussed, with special reference to the pentode design, which enables heavy currents to be controlled from high-impedance grid circuits. The arc voltage drop varies considerably with filling pressure and current. If the current-carrying capacity of the arc path is exceeded, the arc may suddenly go out, and undesirable h.v. surges may result. A method of measuring deionization time is described. Results are given showing how this time varies according to the nature of the filling gas and its pressure. See also 3168 of 1949 (Jones et al.).

621.385.5 : 621.396.622.7 : 621.396.619.13 505

The 'ϕ-Detector', A Detector Valve for Frequency Modulation.—J. L. H. Jonker & A. J. W. M. van Overbeek. (*Philips tech. Rev.*, July 1949, Vol. 11, No. 1, pp. 1–11.) Description of a new valve with 7 grids, which can replace several circuits and valves needed in other detection systems. The second, fourth and sixth grids are screen grids and the seventh is a suppressor grid. The third and fifth grids are control grids, to each of which an output voltage is applied from an i.f. transformer. The r.m.s. value of these voltages must be at least 8 V. The mean value of the anode current is a function of the phase shift ϕ between the two control voltages. ϕ is a function of the frequency deviation. Both these functions are approximately linear when ϕ has a sweep between 60° and 120°. The amplitude of the anode current is independent of the magnitude of the control voltages, provided these exceed 8 V. The valve thus acts as a limiter; this renders certain sources of noise and of distortion harmless. No inertia other than electron inertia is involved, so that short impulsive interference bursts are also limited. The 'ϕ-detector' has an a.f. output voltage of 20–25 V. No i.f. transformer is needed. The first grid can be used for blocking the cathode current if the control voltages are not large enough, thus suppressing interchannel noise. Type EQ80 is made in the same way as Rimlock valves but has a 'Noval' base with 9 pins. The mean anode current is 0.25 mA.

621.385.832 506

Cathode-Ray Tubes for the Study of Rapid Phenomena.—P. Patriarche & P. Bonvalot. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 525–531.) Improvement of the performance of c.r. tubes by better design of electron guns, by reduction of distortion and aberration of the electron-optical systems, and particularly by the use of post-deflection acceleration, is discussed and optimum conditions for the observation of transient effects are considered.

621.396.615.141.2 507

Methods of Measurement used in the Study of Magnetrons. Results obtained on New Types of Magnetron.—J. Legros & C. Azéma. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 568–576.) Measurements necessary prior to evacuation, such as the determination of the resonance frequencies of the anode block, and measurements of various parameters for completed magnetrons under pulse or c.w. conditions, are outlined. Operating characteristics are given for the following new valves: (*a*) Type MV.201, a miniature valve with glass envelope and overlapping segments giving a mean power of 6 W on wavelengths from 19 to 28 cm, with frequency variation by means of an external circuit. (*b*) Type MCV.81, for c.w. operation with a power of 80 W in the range 8.2–8.8 cm, with mechanical regulation. (*c*) Type MC.1011, for high-power radar pulse operation; the anode block has double strapping on both faces; air-jet cooling is used and the useful power with 1-μs pulses at 500 per sec can reach 1 MW. (*d*) Type MC.231, for c.w. operation on a wavelength of about 23 cm, with a useful power of 1 kW; the anode is water-cooled, but air-cooling will be used on later models.

621.396.615.141.2 508

The Modes of Oscillation for Magnetron Anodes.—C. Azéma. (*Bull. Soc. franç. Élect.*, Oct. 1949, Vol. 9, No. 97, pp. 559–567.) Two methods are available for determining the resonance frequency for cavity magnetrons, one based on calculations of e.m. fields by the

methods of mathematical physics, the other based on the concept of equivalent circuits, to which ordinary circuit technique can be applied. Both methods are used to obtain information as to the characteristics of the frequency spectrum for magnetrons with different types of strapping, and for rising-sun magnetrons. Three methods of mode isolation that have been tried by the Compagnie générale de T.S.F., involving the use of rounded fins, decoupled cavities or unequal cavities, are briefly mentioned.

621.396.615.141.2 509

Modes in Interdigital Magnetrons.—J. F. Hull & L. W. Greenwald. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1258–1263.) A set of design equations is derived, by means of which the resonance wavelength and external Q can be calculated for each mode of these magnetrons. Results thus obtained are compared with experimental results. See also 908 of 1949 (Hull & Randals).

621.396.615.142 : 621.385.029.63/.64 510

Electron Bunching in a Velocity-Modulation Valve by means of a Travelling-Wave Device.—R. Warnecke, W. Kleen, O. Doehler & H. Huber. (*C. R. Acad. Sci., Paris*, 3rd Oct. 1949, Vol. 229, No. 14, pp. 648–649.) The gain of travelling-wave amplifiers of the Kompfner-Pierce type is of the order of magnitude of the ratio of the real part to the imaginary part of the propagation constant of the amplified wave; only the energy corresponding to a small excess of the electron velocity over the travelling-wave velocity can be converted to h.f. e.m. energy. The efficiency of the energy exchange can be improved by means of a system in which the electron beam, modulated in density and velocity by means of a wave applied at the input of a retarding system, is injected into a cavity resonator in the field of which the electron packets give up a large part of their kinetic energy. The h.f. energy transported by the travelling wave to the output end of the retarding system is either absorbed by a matched attenuating element or added, by means of a phase transformer, to that which appears in the resonator. This arrangement can thus be considered as an improvement of the ordinary travelling-wave valve, or of the v.m. valve with two interacting fields. Theory of this method of amplification is summarized. Electron bunching and energy conversion factors as high as those for v.m. valves can be obtained. The method has the advantage of requiring no tuned device, with the properties of an oscillatory circuit, at the input. This results in an essential simplification of the adjustments and avoids the limitation of the pass band which, in klystron amplifiers, involves the high Q of the tuned device. With the new amplifier, in fact, the pass band is that of the output cavity resonator and this band is widened by the operational load.

621.396.645 511

Amplification by Direct Electronic Interaction in Valves without Circuits.—P. Guénard, R. Berterottière & O. Doehler. (*Bull. Soc. franç. Elect.*, Oct. 1949, Vol. 9, No. 97, pp. 543–549.) See 2977 of 1949.

621.396.645 : 537.311.33 : 621.315.59 512

The Theory of *p-n* Junctions in Semiconductors and *p-n* Junction Transistors.—Shockley. (*See* 379.)

621.396.645 : 537.311.33 : 621.315.59 513

Hole Injection in Germanium — Quantitative Studies and Filamentary Transistors.—Shockley, Pearson & Haynes. (*See* 380.)

621.396.645 : 537.311.33 : 621.315.59 514

Some Circuit Aspects of the Transistor.—Ryder & Kircher. (*See* 334.)

621.396.822 515

On the Theory of the Shot Effect.—L. A. Weinstein. (*Zh. tekh. Fiz.*, Sept. 1947, Vol. 17, No. 9, pp. 1045–1050. In Russian.) A mathematical discussion of the shot effect in a cylindrical diode when the current is limited by space charge, taking into account the transit time of electrons. When space charge is present the velocity distribution of emitted electrons must be considered. Further investigations are required to determine the relation between the fluctuations of the velocity distribution and those of the anode current.

621.396.822 516

Suppression of Shot Effect Noise in Triodes and Pentodes.—K. S. Knol & A. Versnel. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 462–464. In English.) Strutt & van der Ziel (3358 of 1948) suggested that spontaneous fluctuations arising from the shot effect could be suppressed by using induced grid noise and capacitive detuning of the input circuit. The connection of a capacitor between the control grid and the anode results in a much larger suppression of noise by detuning than is possible without the capacitor. Noise measurements at frequencies of 43 and 120 Mc/s on a Type-EF50 valve used first as a triode and then as a pentode are discussed. The method of measurement was analogous to that of van der Ziel & Versnel (249 of 1949).

621.396.822 517

Measurements on Total-Emission Conductance at 35 cm and 15 cm Wavelength.—G. Diemer & K. S. Knol. (*Physica, 's Grav.*, July 1949, Vol. 15, Nos. 5/6, pp. 459–462. In English.) An experimental disk-seal diode was used, the cathode/grid spacing being 15 μ at a saturation current of 35 mA and 20 μ at a saturation current of 0.1 mA. The active area was 0.75 mm^2. Measurements were made at cathode temperatures between 1 100°K and 1 350°K. The valve was studied in a coaxial circuit loosely coupled to a modulated standard signal generator and also to a crystal detector and l.f. amplifier; the valve conductance was calculated from Q measurements on this circuit for various saturation currents and anode voltages. Results are shown graphically and discussed; they do not support the 'linear field theory'. See also 3310–3313 of 1949.

621.396.822 : 621.385 518

Shot Effect in Valves and the Limits of Amplification.—K. Fränz. (*FIAT Review of German Science, 1939–1946: Electronics, incl. Fundamental Emission Phenomena*, Part 2, 1948, pp. 21–29. In German.) A general review, with references to 19 German publications. Input-circuit design is considered.

621.385 519

Fundamentals of Radio-Valve Technique. [Book Review]—J. Deketh. Publishers: Cleaver-Hume Press, London, 535 pp., 35s. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, p. 413.) A translation of the first of a series of books on valves originated by Philips, Eindhoven. A fairly elementary treatment suited to the student and with useful references for the engineer.

MISCELLANEOUS

5 + 6'(43) 520

FIAT Review of German Science, 1939-1946: Electronics, incl. Fundamental Emission Phenomena: Part 2. [Book Notice]—G. Goubau & J. Zenneck (Senior Authors). Publishers: Office of Military Government for Germany, Field Information Agencies Technical, British, French, U.S., 1948, 288 pp. In German. Part 1 was noted in 2414 of 1949. For abstracts of papers see various sections.

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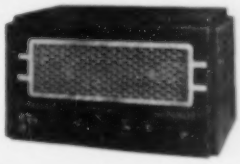

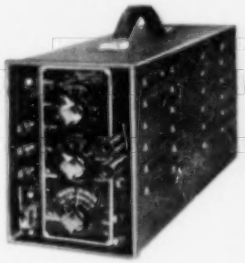

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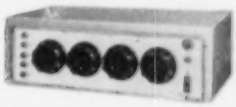

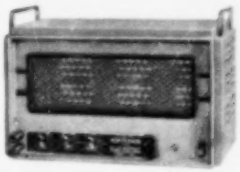

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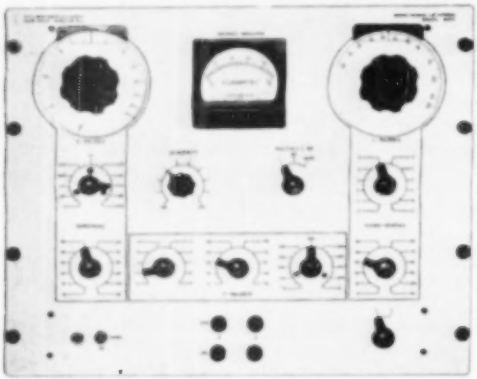

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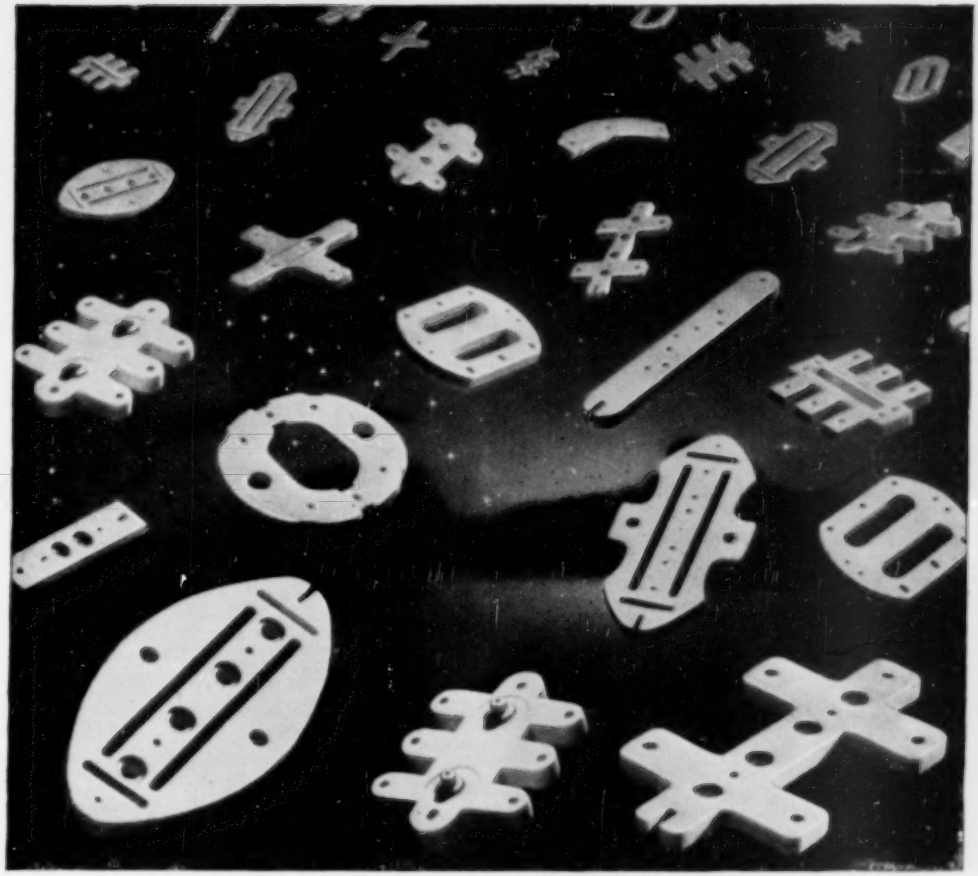

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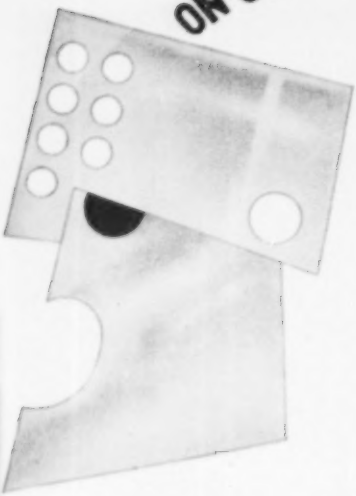

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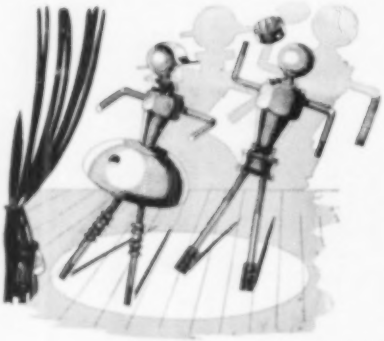

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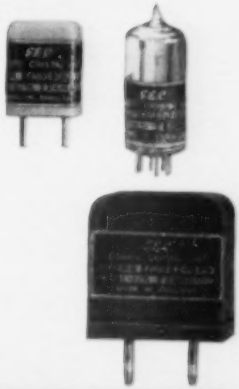

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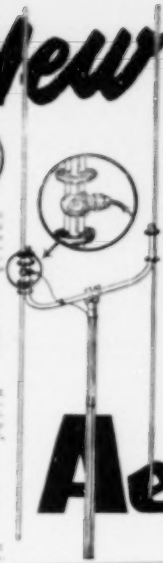

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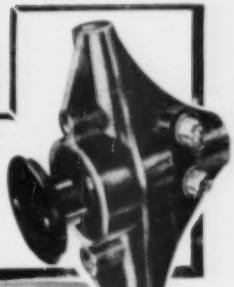

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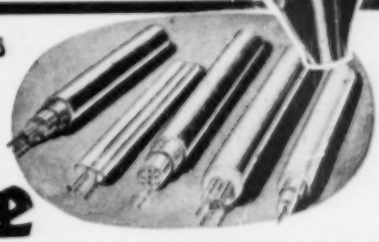

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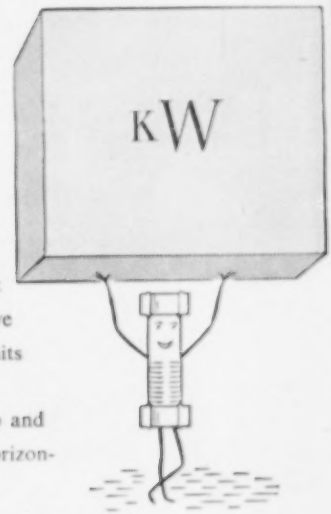

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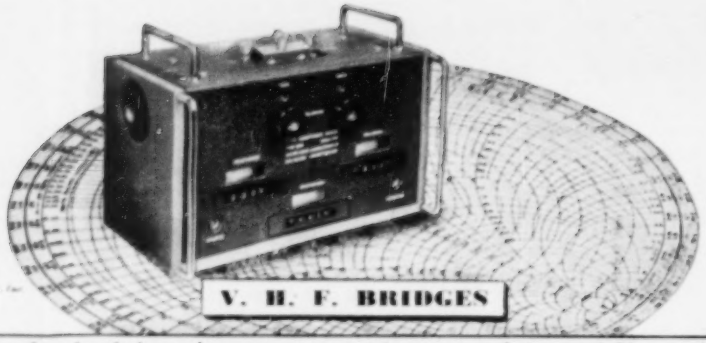

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If the problem is Transformers or Chokes none are more experienced to find the solution than Partridge. Nearly two decades of specialised research and production have prompted leading industrial concerns and Government Experimental Stations constantly to use the unequalled Partridge service. Your need, too, can be adequately met for transformers ranging from 5 VA to 50 KVA—power or A.F. ***Send for catalogue and data sheets.***

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MEET SOME OLD FRIENDS

THERE'S no friend like an old friend — and we have been making some of these friends since 1906: we lose sight of them very quickly, but we often hear about the long and reliable service they are giving. So — in case it is "out of sight — out of mind", we take this opportunity of reminding you of their sterling qualities and wide range.

Rectangular Metal Cans.
Range: 250 to 30,000 V. D.C. *working.*
Capacitance: .01 to 10 *Mfds.*

The Moulded Mica types are also friends of long standing: they have been used in millions since their inception in pre-broadcasting days, and are regarded as the standard of condenser reliability.

Moulded stacked foil Mica.
Range: .00005 *Mfds. at* 350 V. D.C. *working to* .0025 *Mfds. at* 2500 V. D.C. *working.*

T.C.C. Condensers are exclusively specified in the View Master — the Television Set you build at home from standard parts. Constructor Envelopes (Model A, London or Model B, Sutton Coldfield) 5/- each from all Wireless Shops.

THE TELEGRAPH CONDENSER CO. LTD

RADIO DIVISION: NORTH ACTON, LONDON, W.3 Telephone: ACORN 0061

WHY IT PAYS TO USE
Ersin Multicore Solder

Radio Manufacturers and Service Engineers all over the world prefer to use *Ersin Multicore Solder* in their factories and workshops. Even in the U.S.A. where more cored solder is produced than anywhere else in the world, many leading firms insist on British made *Ersin Multicore*, despite the freight charges and import duties which raise its cost above that of the locally made single-cored product. Below are some of the reasons why *Ersin Multicore* has attained such world-wide popularity.

FOR MANUFACTURERS

Ersin Multicore Solder is the only solder containing 3 cores of non-corrosive Ersin Flux. Ersin is a high grade rosin which has been subjected to a complex chemical process to increase its fluxing action, whilst still retaining the non-corrosive properties. Ersin Flux makes precision soldering quicker and more economical—it not only prevents oxidation during soldering, but actually cleans the surface to be soldered, removing any oxide from the metal.

Three cores of flux ensure flux continuity throughout the length of the solder wire. There are no lengths without flux—that means no wasted solder, no wasted time, and freedom from 'dry' or H.R. joints.

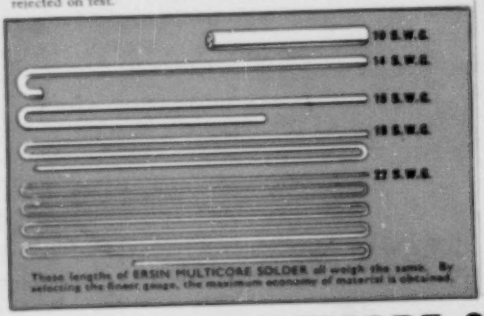
The correct proportions of flux to solder are always assured—no extra flux is required. Three cores of flux provide thinner solder walls, giving instantaneous melting—an important factor in speeding up repeat soldering operations.

Soldered joints made with Ersin Flux do not corrode even after prolonged exposure to any degree of humidity. This has been tested under climatic conditions ranging from the Arctic to the Tropics.

Only the finest virgin tin and lead are used in the manufacture of *Ersin Multicore*, giving complete freedom from impurities.

CUTS PRODUCTION COSTS

The economies effected by using *Ersin Multicore Solder* play an important part in cutting production costs and keeping down the price of radio equipment. You get more joints per lb. of *Ersin Multicore*—there is no waste. Soldering with *Ersin Multicore* is quicker, too, and every joint is a perfect electrical connection. This saves time in your inspection department—fewer sets are rejected on test.



These lengths of ERSIN MULTICORE SOLDER all weigh the same. By selecting the finest gauge, the maximum economy of material is obtained.

FOR FACTORY USE

Ersin Multicore Solder is made as standard for factory use in 5 alloys and 9 gauges, and is supplied on nominal 7 lb. reels. Other Alloys and Gauges can be supplied to special order. Bulk prices on application.

ALLOYS. This table shows the Melting Points of the three principal Alloys used in Electronic and Telecommunication industries.

Alloy	Colour Code	Sol.	Liq.	Bis. Temp.	USES
60/40	Red	183°C	190°C	230°C	High quality work
45/55	Crimson Buff	183°C	227°C	267°C	Hand soldering radio and electrical
40/60	Green	183°C	238°C	278°C	

GAUGES. Approximate number of feet per lb.

Standard Wire Gauge	Diam. in inches	Diam. in M.ms.	ALLOY 60/40	ALLOY 45/55	ALLOY 40/60
10	0.128	3.251	25.2	23.5	23.0
12	0.104	2.642	38.1	35.2	34.9
13	0.092	2.337	48.7	45.3	44.5
14	0.080	2.032	64.4	59.2	58.6
16	0.064	1.626	100.5	94.3	92.1
18	0.048	1.219	178.5	167.8	163.5
19	0.040	1.016	257.5	240.4	235.5
20	0.036	0.914	318.0	302.5	291.0
22	0.028	0.711	526.0	492.0	481.0

FLUX FACTS ABOUT ERSIN MULTICORE SOLDER

Ersin Multicore Solder is made in a wide range of flux speeds and percentages. The following information, which should be studied in conjunction with the tables of alloys and gauges, will help you to select the most suitable type for your particular job.

1 ERSIN MULTICORE SOLDER with N flux. This is a development of the original ERSIN MULTICORE SOLDER, having a slightly faster flux speed than the original flux (Standard flux 5E). It has thus been possible to reduce the flux percentage to approximately 2.2%. This grade is normally supplied unless an order specifies otherwise.

2 ERSIN MULTICORE SOLDER with Standard flux 5E has a flux percentage of 3.4%.
ERSIN MULTICORE SOLDER types 1 and 2 are supplied in all Tin/Lead alloys, and other special alloys, and in most gauges between 10 and 22 S.W.G. They can be supplied in other diameters to special order. These solders comply with all Government and Ministry specifications, including M.A.P. D.T.D. 599 and are widely used in the Radio, Television, Telephone and Electronic industries.

3 ERSIN MULTICORE SOLDER with L flux.
The flux in this solder is similar to 1 and 2, but contains a higher percentage of the activating agent, making it specially suitable for use with the high-speed machines employed in the Lamp and Battery industries. The flux complies with the proposed new B.S.I. Cored Solder specification. The flux percentage is 3.4%.

4 ERSIN MULTICORE SOLDER with 2L flux is similar to type 3 but has a flux percentage of 2.2%.
Types 3 and 4 are supplied as standard only in 20/80 and 40/60 alloys. Other alloys can be supplied to special order.

Write for comprehensive technical information and free samples of any specifications.

MULTICORE SOLDERS LTD.

MELLIER HOUSE, ALBEMARLE STREET, LONDON, W.1 • Telephone: REGent 1411